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FINAL DEVELOPMENT REPORT
FOR
UNATTENDED ELECTRONIC FERRET
RECONNAISSANCE SET AN/DLD-2(XA-1)

by

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WRIGHT AIR DEVELOPMENT CENTER
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ABSTRACT

1 This final report describes the development of Unattended Ferret Reconnaissance Set AN/DLD-2(XA-1). This ferret is a transistorized, wide-open, automatic intelligence-gathering system operating in the 30 to 1000 Mc range. The following parameters of intercepted radars are measured: frequency, direction of arrival, PRF, and pulse width. The measured parameters are provided in digital form and can be either stored on magnetic tape or transmitted on a digital data link (or both). A feasibility demonstration model of two full sub-bands (60 to 120 Mc and 480 to 1000 Mc) was constructed and demonstrated.

2 A complete review of the system organization is given. The detailed development program, comprising development of the following sub-units of the system and their performance characteristics, is discussed: antennas, television-interference elimination, direction analyzer, frequency analyzer, pulse-width analyzer, airborne computer and PRF analyzer, read-out device, power supply, and special test equipment. The detailed outline of the feasibility demonstration of the breadboard system is described, including detailed measurements of system performance. Reliability predictions and the reliability assurance program are also discussed.

3 The following results of measurements made on the feasibility demonstration model, including aircraft environment (B-47 model), showed:

Frequency	+2% average accuracy, ±4% maximum error
Direction	+12 degrees maximum error over each 120-degree azimuth sector

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PRF +2% over the 20 cps to 10 kc
Range

Pulse Width +5% or 1 μ sec over the 1 to
100 μ sec range

The basic conclusion reached is that the AN/DLD-2 equipment is feasible.

- 4 Finally, recommendations for future AN/DLD-2 equipments are made. These recommendations include specific areas for techniques improvement, such as antennas, line-probing techniques, television-interference elimination, and smaller and lower-power memories.

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PART I

A. PURPOSE

- 5 The purpose of this project is to conduct a study and investigation leading to the development and fabrication of transistorized Unattended Electronic Ferret Reconnaissance Set AN/DLD-2(XA-1). This study and investigation is to advance the state of the art in unattended ferret reconnaissance in the 30 to 1000 Mc range. The reconnaissance equipment will provide means for intercepting, direction finding, and analyzing signals in the band and will process and store frequency- and direction-finding information for subsequent rapid analysis and dissemination.

B. GENERAL FACTUAL DATA1. IDENTIFICATION OF TECHNICAL PERSONNEL

- 6 In the development of this equipment, the following engineering personnel participated to the extent indicated.

		Hours
R. A. Angevine	Engineer	639
W. R. Aylward	Group Leader	1059
J. G. Birosh	Engineer	9
A. Boecker	Engineer	2586
H. Briggs	Engineer	126
R. Burfeind	Engineer	22
J. F. Craib	Consultant	166
G. F. Dery	Engineer	77
L. K. DeSize	Engineer	305
G. M. Froehlig	Engineer	1864
W. E. Fromm	Department Head	138
E. C. Heith	Engineer	671
R. Jacobl	Engineer	248
W. J. Keane	Engineer	321
J. W. Kearney	Section Head	1043
S. H. Klug	Engineer	1491
R. F. Koch	Assistant Proj. Eng.	2659

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		Hours
L. LoSasso	Engineer	941
J. A. Lovell	Engineer	8
R. V. Lowman	Project Engineer	2705
Y. J. Lubkin	Engineer	749
R. G. Malech	Engineer	13
J. F. McDonald	Engineer	1347
J. A. McDonough	Section Head	109
C. J. Meiselbach	Engineer	975
V. V. Milukas	Engineer	2057
F. J. Mueller	Engineer	1928
W. M. Mulqueen	Engineer	1482
K. S. Packard	Consultant	4
W. P. Peyser	Engineer	122
D. W. Pride	Engineer	1554
F. Rappolt	Engineer	242
B. Sacks	Engineer	705
I. M. Saffitz	Engineer	2745
H. Salz	Engineer	1470
E. W. Sard	Engineer	895
W. Satre	Engineer	16
R. J. Schniebolck	Engineer	491
L. W. Schrader	Engineer	1844
B. Segall	Chemist	20
R. F. Simons	Department Head	14
E. Sion	Engineer	178
M. R. Van Dusen	Engineer	2056
K. Weinberg	Engineer	547
D. R. Weller	Section Head	60
A. L. Williams	Engineer	2105
W. H. Yale	Engineer	1333

2. PATENTS

7 The patentability of the following contributions to the electronic art conceived in the course of this project is being investigated:

Minimum-amplitude selector,
Direction-analyzer ambel,
Television-interference suppressor,
Quantized phase-comparison direction-finding,
Assembly of probed lines fed by directional couplers,
Hybrid tree,
R-F detector mounts,

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Distributed-constant hybrid ring,
Lumped-constant hybrid ring,
Distributed-constant directional coupler,
UHF filter bank,
Tester for video-detector crystals.

3. FORMULAS

8 The major formulas giving mathematical descriptions of equipment functions are shown in the sections on the direction analyzer, the frequency analyzer, the computer and PRF analyzer, and special tests and test equipment. The formulas are given with the theory of operation of these parts of the equipment.

4. MEASUREMENT PROCEDURES

9 The following specialized measurement procedures have been used on this project:

Three-antenna method of measuring antenna gain (see paragraph 64),
Use of hybrid rings for r-f phase measurements (see paragraph 65),
Superheterodyne detection of standing waves on a slotted line (see paragraph 56),
Special receiver for evaluation of interfering television signals (see paragraph 71),
Measurement of amplitude and phase tracking of pairs of r-f amplifiers (for distributed amplifiers, see paragraph 104; for traveling-wave tubes, see paragraph 107),
Measurement of recovery time of video amplifiers (see paragraph 113),
System tests of the feasibility-demonstration model (see paragraphs 302 through 365).

10 We have built the following specialized test units:
Transistor tester (see paragraphs 265 through 267),
Test-problem generator for computer (see paragraphs 293 through 297),

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Pulse-train generator (see paragraphs 290 through 292),
Tester for video-detector crystals (see paragraphs 281 through 289).

5. REFERENCES

The following references are cited in this report:

1. W. R. Aylward et al., "Interim Development Report for Unattended Ferret Reconnaissance Set AN/DLD-2(XA-1)," Report No. 3642-I-1, Airborne Instruments Laboratory, Inc., October 1956.
2. J. W. Kearney et al., "Interim Engineering Report on Electronic Ferret Reconnaissance System AN/DLD-1," Report No. 2651-I-1, Airborne Instruments Laboratory, Inc., January 1955.
3. A. Boecker et al., "Interim Development Report for Unattended Electronic Ferret Reconnaissance Set AN/DLD-2(XA-1)," Report No. 3642-I-2, Airborne Instruments Laboratory, Inc., December 1956.
4. R. F. Koch et al., "Interim Development Report for Unattended Electronic Ferret Reconnaissance Set AN/DLD-2(XA-1)," Report No. 3642-I-3, Airborne Instruments Laboratory, Inc., 1 March 1957.
5. S. H. Klug et al., "Interim Development Report for Unattended Electronic Ferret Reconnaissance Set AN/DLD-2(XA-1)," Report No. 3642-I-4, Airborne Instruments Laboratory, Inc., 3 June 1957.
6. S. H. Klug et al., "Interim Development Report for Unattended Electronic Ferret Reconnaissance Set AN/DLD-2(XA-1)," Report No. 3642-I-5, Airborne Instruments Laboratory, Inc., 3 September 1957.
7. "Reliability Stress Analysis for Electronic Equipment," TR-1100, Radio Corporation of America, Defense Electronics Products, Camden, New Jersey, November 1956.

6. MATERIALS

There is nothing to report under this heading.

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C. DETAIL FACTUAL DATA1. OVER-ALL SYSTEM

- 13 Unattended Electronic Ferret Reconnaissance Set AN/DLD-2(XA-1) is an automatic intelligence-gathering type of radar countermeasures intercept system that is capable of high rates of data acquisition. It is based upon the use of (1) multichannel receivers to achieve nonscanning frequency analysis, (2) fixed-position antennas with wide-open receivers to achieve nonscanning (instantaneous) direction finding, and (3) post-receiver sorting, storing, and recording of data in digital form. The following parameters of intercepted radars are measured: frequency, direction of arrival, pulse repetition frequency (PRF), and pulse width. The system includes a computer that performs important signal analyses prior to recording. In the basic design of the equipment, particular emphasis is placed upon conditions of high signal-traffic densities and wide dynamic range of signal strengths.

- 14 Figure 1 is a block diagram of the AN/DLD-2(XA-1), which covers the 30 to 1000 Mc range. Five sets of antennas are used, each set covering about a 2:1 frequency sub-band. Each antenna set responds to vertically and horizontally polarized signals, and the composite pattern of each set is roughly omnidirectional in the hemisphere below the aircraft. These antenna characteristics satisfy the requirements of the frequency analyzer and the direction analyzer. The frequency ranges covered by the five sets of antennas are as follows:

Sub-Band No.	Frequency (Mc)
1	30 to 60
2	60 to 120
3	120 to 240
4	240 to 480
5	480 to 1000

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15 The direction analyzer uses a phase-comparison method of direction finding. Basically, this method uses two antennas and means for comparing their outputs. When a signal is received, each antenna samples the arriving waves. The difference in time of arrival of the samples is a function of the angle between the center-to-center line of the antennas and the direction from which the received signal arrives. Thus, when the antenna outputs are appropriately compared, the angle of arrival of the signal can be determined. Since the same difference in time of arrival occurs for signals emanating from the right or the left hemisphere, a mirror ambiguity results and a separate left-right indicator is used to resolve it.

16 To overcome the limited azimuthal coverage of flush-mounted antennas, the set of antennas for each sub-band of an operational system consists of two pairs of antennas--one pair mounted on the left side of the aircraft and the other on the right side. Signals from the two forward antennas are combined, as are the outputs from the two aft antennas; the combined signals are fed to the two inputs of the phase-comparison section of the direction analyzer. The analyzer operates upon these signals as though they were derived from a single pair of antennas having horizon-to-horizon coverage below the aircraft. Thus, the absolute value of the angle of arrival of signals is obtained. The sign of the angle is determined by a left-right indicator that is part of the direction analyzer. This indicator derives inputs from left and right antennas separately, measures their relative signal strength, and resolves the ambiguity accordingly.

17 In sub-band 1, only left-right direction indications are produced. Consequently, the antenna set for this sub-band includes only one at the right and one at the left.

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18 The phase-comparison equipment performs its function in a quantized manner. Each angle of arrival is determined as lying within one of 15 direction channels. Eleven channels are included within the two 120-degree sectors that point symmetrically to left and right of the aircraft. The remaining four channels cover the 60-degree forward and aft sectors; signals arriving from these directions are not further processed by the ferret system, because direction cuts approaching parallelism with the aircraft heading are of little use in station location. The output from the direction analyzer is a 5-bit word, of which 4 bits indicate channel numbers and one gives left-right information.

19 R-F signals, amplified in the direction analyzer, are fed to the frequency analyzer, where the frequency of each incoming signal is determined. This frequency analysis is performed by channelizing r-f filters, each of which drives a crystal-video receiver. The 30 to 1000 Mc range is divided into 45 channels by the r-f filters. Because these filters do not have perfectly rectangular pass bands, strong signals can produce responses in two or even three adjacent channels. To take care of such cases, the outputs from the channel receivers are processed by an ambiguity eliminator, or ambel. The logic of the ambel is such that if two adjacent responses are produced, the ambel inserts an interpolate channel between the two filter channels. This interpolation indicates that the frequency of the received signal is about midway between the centers of the responding channels. If three adjacent responses are produced, the ambel selects the response on the middle channel. By introducing these interpolate channels, the ambel increases the number of frequency channels to 89. The output from the frequency analyzer is a 7-bit word that identifies the channel in which each signal is received.

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- 20 Gating circuits demand time coincidence of direction and frequency signals for admission of the signals to the computer. This coincidence requirement acts as a noise-reduction device; noise that produces pulses in only one analyzer is barred from the computer because the time-coincidence requirement is not then satisfied.
- 21 The received pulse, standardized in amplitude, is applied to the pulse-width analyzer by the direction analyzer. The leading edge of the pulse starts a start-stop oscillator; the pulses generated from this oscillator at a time rate are applied to a binary counter. The trailing edge of the received pulse stops the start-stop oscillator. The counter thus counts 1- μ sec intervals during the time that the received pulse is present. The output from the pulse-width analyzer is a 7-bit word giving the duration of the received pulse in microseconds--between 1 and 126.
- 22 PRF is determined by measuring its reciprocal--recurrence interval or pulse period--in the computer. The measurement is made by counting the number of pulses generated by a crystal-controlled oscillator in the interval between pulses of a given signal; the rate of the oscillator is determined by the precision of measurement required. The counting is performed in a quasi-logarithmic manner to reduce the number of bits used to measure long intervals. Quasi-logarithmic counting is obtained by beginning the counting process at a high rate, and reducing the rate, in successive steps, as the total count increases. The resultant PRF word consists of 10 bits.
- 23 Frequency, direction, and pulse-width information are fed to the computer, which sorts the input parameters pertaining to a given radar intercept and internally measures the PRF parameter. The computer also verifies intercepts (it rejects intercepts that do not recur at a rate greater

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than 20 pps) and provides temporary storage of the intercept prior to recording. The computer consists of two basic parts: (1) temporary storage units (called trunks), and (2) supervisory circuits that control the flow of signals through the computer. The trunks store data and measure PRF. The PRF counters serve both the counting and storing functions.

- 24 There are five trunks in our present design; thus, five intercepts can be stored simultaneously, and provision is made that no two intercepts are alike in frequency and direction. Therefore, the multi-trunk system performs a redundancy-reduction function that ensures high probability of intercept in areas of high signal-traffic densities. Each trunk also performs verification, which is another noise-reduction function in addition to that performed by the coincidence requirement for admission of frequency and direction data to the computer. Verification is obtained in each trunk by demanding that an intercept be repeated at least twice within an interval of 0.1 second (two pulse periods at a 20-pps rate) after it is first stored in the trunk. If this minimum repetition requirement is not met, the trunk is cleared.
- 25 The supervisory circuits consist of two controls and a central clock for the PRF counter. One set of controls is started when data pulses are fed to the computer by the frequency and direction analyzers. This set interrogates the trunks and routes new frequency and direction data to the lowest-numbered trunk not already in use; the question of whether the data are new is resolved by comparisons made in each trunk under the control of the supervisory circuits. The other set of controls, which is repeated in each trunk because of its simplicity, routes pulse-width data appropriately.
- 26 The read-out unit sequences through the computer trunks, reads out the stored information, and transfers the information to a magnetic tape recorder. After a trunk is

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read out, it is reset to receive new information. The read-out unit also sequences periodically to a navigational digitizer [not included in the AN/DLD-2(XA-1) system], to record on the tape pertinent navigational data, such as time, latitude, longitude, altitude, and aircraft heading. The read-out program provides for two sequences through the computer and one through the navigation digitizer each second, thus establishing the rate of intercept information flow as 10 per second.

27 The magnetic tape recorder is a 16-track unit that records data in a format suitable for analysis by the AN/GLA-7 (a ground-based data-handling system). The data capacity of a single reel of tape is adequate for at least a 10-hour mission, during which as many as 360,000 intercepts may be recorded.

28 The AN/DLD-2(XA-1) is designed to have maximum versatility in its operational use. Each major unit consists of a set of building blocks (functional sub-units) in an Erector-set manner. Thus, portions of the system could be used as the need arises, without requiring further development. For example, in the fighter type of reconnaissance aircraft where space is at a premium, a partial system could be installed with limited frequency coverage. Unitized modular construction, with plug-in subminiature chassis, would be used throughout.

29 The principal performance objectives of the AN/DLD-2(XA-1) are as follows:

R-F range	30 to 1000 Mc
Sensitivity	Equivalent to -40 dbm with isotropic antenna
Dynamic range	60 db
Frequency resolution	+2 percent average, +4 percent minimum
D-F accuracy	+6 degrees average, +12 degrees minimum

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Azimuth coverage	Two 120-degree sectors symmetrically abeam of the aircraft
PRF range	20 to 10,000 pps
Pulse-width range	1 to 126 μ sec
PRF accuracy	± 2 percent
Pulse-width accuracy	± 5 percent + 1 μ sec
Rate of intercept	5 or more different radars per second

30 Where possible, the designs of transistorized units for the AN/DLD-2(XA-1) were made compatible with the units performing similar functions in the AN/DLD-1. The AN/DLD-1(XA-1) uses vacuum tubes; when a transistorized AN/DLD-1 is built, the task will be greatly simplified through the availability of circuit designs that can be adapted with little or no changes from the AN/DLD-2(XA-1).

2. ALTERNATIVE SYSTEM POSSIBILITIESa. FREQUENCY-SCANNING RECEIVER

31 At the beginning of this project, we studied the relative merits of two systems of ferret reception (reference 1). These systems are the wide-open type pioneered by AIL and used in the AN/DLD-1 (reference 2), and a frequency-scanning type embodying new features suitable for an unattended equipment. The two systems share a most important operational concept--both reduce instantaneous received signal density by introducing a scanning process. Otherwise, the receiver and the airborne data-handling circuits would have to be impractically large and complex to accept the many signals present in high-density environments. The scanning process used by the wide-open system is the antenna-scanning of the target radars, since receiver sensitivity has the characteristic that targets are seen only when their antennas look at the receiver's antennas. The frequency-scanning system uses the scanning process for which it is named. Since the receiver sensitivity is sufficient to

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pick up the background radiation of the target radars, the effects of the scanning of the radar antenna are eliminated.

32 In the course of this study, we summarized what we consider to be the relative advantages of the two systems (see Table I). Figure 2 shows the proposed frequency-scanning receiver.

33 The frequency-scanning system shown in Figure 2 uses a superheterodyne direction receiver and frequency-determining receiver. A single set of scanning local oscillators is used for both receivers. Consequently, the frequency at which the direction receiver is operative at any instant is the same as that of the frequency receiver. In other respects, the frequency-scanning direction receiver resembles the wide-open direction receiver.

34 The frequency-determining receiver consists of five sub-bands (as with the wide-open system), each having its own receiving equipment. The local oscillators for all sub-bands are driven continuously and simultaneously from a common drive. Digitized frequency-channel information will be derived from this common drive in a code wheel or similar device. The most logical unit for digitizing is one i-f bandwidth. This digitized signal and a pulse that indicates the presence of signal in one of the sub-bands provides complete frequency information on the incoming signal. The detected pulse from the sub-band receiver is used for pulse-width measurement.

35 Limits can be set on the selection of parameters for the receiver. Maximum frequency resolution is determined by i-f bandwidth and r-f pulse spectrum. Maximum scan rate is determined by minimum PRF and i-f bandwidth. Two pulse intervals are required to measure PRF. The minimum number of sub-bands is limited to about three because of antenna and r-f circuit considerations.

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TABLE I
RELATIVE MERITS OF WIDE-OPEN AND FREQUENCY-SCANNING SYSTEMS IN
30 TO 1000 Mc RANGE

Consideration	Wide-Open System	Frequency-Scanning System
Range		
a. Pulse signals	Probably adequate for main lobe of transmitting antenna	Adequate even for background radiation of transmitting antenna
b. C-W signals	Probably inadequate	Probably adequate
Frequency resolution	Improvement requires increasing number of channel filters and auxiliary equipment	Improvement requires change in i-f bandwidth and minor changes in computer
Probability of intercept		
a. For full-circle scan of antenna of target radar	Good	Good
b. Sector scan	Fair	Good
c. Burst transmission if for full-circle scan	Good	Uncertain; depends upon relative periods of burst and receiver's frequency scan
d. Frequency diversity radars	Poor to fair	Fair to good
e. C-W signals	Poor	Good
f. D-F information	Good in light traffic; fair in heavy traffic	Good
Heavy traffic	Poor; requires very short recovery times and many storage trunks	Good; frequency sweep reduces traffic per unit time

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TABLE I (cont)

Consideration	Wide-Open System	Frequency-Scanning System
Freedom from interference due to:		
a. Strong signals with broad spectra	Good	Fair
b. Strong cw	May jam direction receiver	Will jam only one spot in frequency band
c. Weak cw	Good	May jam one spot in frequency band
Direction-finding range	Poor; requires r-f amplification	Good
Adaptability to c-w analysis	Very difficult	Practical
Frequency-band overlap	Compensated by frequency ambel	Requires accurate setting of band edges; produces redundant and possibly erroneous information on signals in overlap range
Identification of multi-frequency radars		
a. Synchronous	Can provide for recognition and recording as such	Will be received as individual signals; ground computer may possibly recognize this type by relationship of frequencies
b. Frequency diversity	Will not recognize as such	Will be received as individual signals; ground computer may possibly recognize this type by relationship of frequencies

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TABLE I (cont)

Consideration	Wide-Open System	Frequency-Scanning System
Signal-parameter accuracy		
a. Frequency	Good	Excellent
b. Direction	Depends upon principle used; limited by low sensitivity	Depends upon principle used; not limited by sensitivity
c. Polarization	Impractical because of limitations of airborne antennas	Impractical because of limitations of airborne antennas
d. Pulse width	Good	Fair to good, depending upon tuning
e. PRF	Good	Good
f. Signal strength	Fair	None, unless provision is made to hold receiver on signal to be measured
g. Scan rate	Fair	Same as signal strength
Local radar interference	Receivers must be blanked each time local radar fires	Blanking required only when receiver is tuned to local radar
Transistorizing	All circuits except r-f amplifiers	All circuits except local oscillators
Maintenance difficulties	R-F portions easier, computer more difficult	R-F portions more difficult, computer easier
Reliability	Many parallel signal paths reduce likelihood of complete system failure	Few parallel signal paths increase likelihood of complete system failure

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TABLE I (cont)

Consideration	Wide-Open System	Frequency-Scanning System
Equipment size and weight	Somewhat larger and heavier because of larger memory requirements	Somewhat smaller and lighter because of smaller memory requirements
Power requirements	Approximately equal	Approximately equal
Development time and money	Approximately equal	Approximately equal

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36

The total frequency scan time required to cover the full 30 to 1000 Mc range must be comparable with the lowest radar scan time to ensure high probability of intercept on burst transmissions. This affects the choice of frequency resolution and frequency scan rate. For ease and accuracy of construction of the digitizer, it is desirable to make the r-f bandwidth, the i-f bandwidth, and the scan rate in one sub-band logarithmically related to those in the other sub-bands.

37

One possible set of parameters for such a receiver is as follows:

Sub-bands	30 to 60 Mc, 60 to 120 Mc, 120 to 240 Mc, 240 to 480 Mc, and 480 to 1000 Mc.
I-F bandwidths	0.3 Mc, 0.6 Mc, 1.2 Mc, 2.4 Mc, and 4.8 Mc
Scan Rate	3 Mc per second, 6 Mc per second, 12 Mc per second, 24 Mc per second, and 48 Mc per second.
Total scan time	10 seconds

38

The final selection of parameters would be dictated by the end use of the equipment. If higher probability of intercept for burst transmissions is required, for example, it can be obtained by using more sub-bands or by a sacrifice of frequency resolution.

39

As shown in Table I, the advantages and disadvantages of wide-open and frequency-scanning systems are nearly equal, from a technical viewpoint. There is also the important consideration, however, that use of the wide-open technique in the AN/DLD-2 provides a continuity of system philosophy with the AN/DLD-1. On this basis, it was decided to make the AN/DLD-2 a wide-open receiver (see reference 3).

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b. AMPLITUDE-COMPARISON DIRECTION FINDER

40 At high frequencies, such as those covered by the AN/DLD-1, it is practicable to use a multiplicity of high-gain antennas, each covering a narrow sector, for direction-finding. However, in the 30 to 1000 Mc range, antennas of sufficient gain for such an application are far too large for installation in an aircraft. Consequently, at the beginning of this project, we investigated other direction-finding techniques. Two techniques were considered in detail--phase comparison and amplitude comparison. The Martin Company has developed for the AN/ALD-1 a non-instantaneous phase-comparison direction finder, which utilizes a pair of motor-driven phase shifters to determine direction.

41 The amplitude-comparison direction finder uses a small number of antennas with approximately cosine radiation patterns. Melpar's low-frequency direction system for the AN/ALD-4 uses crossed-dipole antennas. These antennas are horizontally polarized and have filled-in dipole radiation patterns. Since the dipole patterns are crossed, the relative amplitudes of received signals on the two antennas will depend upon the angle-of-arrival of the signal. Suitable circuits measure the relative amplitudes, and the resultant data are eventually reduced to angle-of-arrival data. Because the crossed-dipole antenna configuration has both 90-degree and 180-degree ambiguities inherent in the measurement, Melpar has added a third antenna and suitable circuitry to eliminate the 90-degree ambiguity. The 180-degree ambiguity is resolved when the data are plotted on the ground.

42 The Naval Research Laboratory (NRL) amplitude-comparison system, developed after the Melpar system, uses spiral antennas facing each of four quadrants. Spiral antennas are bipolar; however, size requirements prevent their use at lower frequencies (below 300 Mc). The NRL system has no inherent ambiguities in direction determination.

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43 An important advantage of amplitude-comparison systems over phase-comparison systems is that they have less inherent error due to the depression angle of the target line with respect to the horizontal.

44 On the other hand, this type of phase-comparison system (described in paragraphs 85 to 91) has these advantages over an amplitude comparison system:

1. The antennas can be flush-mounted.
2. The direction data are channelized on distinct buses, and are coded directly as angle-of-arrival as in the AN/DLD-1. Hence, a minimum memory capacity is required in the computer, and radar data redundancy can be removed during temporary storage.
3. Phase-comparison direction finders operate in the null region, which is the region of maximum slope. Consequently, this technique has inherently higher resolution.

3. ANTENNASa. REQUIREMENTS

45 The following antenna specifications are based upon the aircraft environment wherein the antennas are to be used and the requirements of the phase-comparison direction finder. The antennas must (1) have a half-power beamwidth of 120 degrees, (2) be operable over a 2:1 frequency bandwidth, (3) be capable of being flush-mounted, (4) be responsive to both vertically and horizontally polarized signals, (5) be sufficiently small to mount an antenna pair one-half wavelength apart at the highest frequency, (6) have a maximum coupling of about -30 db between a pair of antennas mounted one-half wavelength apart, (7) track in amplitude to ± 2 db (between elements of a pair) over the useful 120-degree sector of their patterns, and (8) track in phase to ± 3 degrees (between elements of a pair) over the useful 120-degree sector.

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46 The development of antennas was divided into two parallel investigations. One investigation was directed toward the development of antennas for the 120 to 1000 Mc sub-bands. In this range, spiral antennas gave the most promise of meeting the specifications, and the effort was concentrated on this type. The second investigation was directed toward the development of antennas for the 30 to 120 Mc sub-bands. In this range, loaded spirals, slot-turnstiles, and tilted-slot antennas were considered. The efforts in the two frequency ranges will be described separately.

b. UNSATISFACTORY APPROACHES

47 During the course of this project, several antenna developments demonstrated limitations of a serious nature. These developments are briefly described here, for the sake of completeness. Greater detail will be found in the interim reports, particularly in references 4 and 5.

48 Unloaded spiral antennas are excessively large in sub-bands 1 and 2. We investigated loading by means of (1) high-dielectric material, and (2) lumped and distributed reactances. Loading with dielectric material greatly increased the weight with only a small attendant size reduction. In this study, we proved that the spiral antenna must include an appreciable loss before the antenna pattern is smooth. With no loss, a reflection from the far end causes a standing wave on the antenna that results in an antenna-pattern break-up. Attempts were made to lump-load the spiral with reactance to reduce the size. Our investigation of this type of loading was limited, and no appreciable size reduction was accomplished. We believe that appreciable size reduction can be realized with little weight increase, provided that sufficient time and effort can be devoted to the investigation.

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49 Slot-turnstile antennas were also considered for the lowest bands. These antennas consist of a pair of slots in space quadrature. When each slot is fed in phase quadrature with respect to the other, the resultant radiation pattern is circularly polarized. Providing a 90-degree phase shift over an octave band is a difficult problem. Since the development of 3-db lumped-constant directional couplers occurred at a later date, this application was not available. Since the slot-turnstile is inherently more complex than a tilted slot and its characteristics are no better for this application, the slot-turnstile development was halted in favor of a tilted slot.

c. ANTENNAS FOR FEASIBILITY MODEL

(1) SUB-BAND 2

50 A cavity-backed slot antenna that is tilted with respect to the vertical provides a simple bipolar antenna configuration. An 18-inch slot antenna scaled 8:1 was designed and used with the 8:1 scale model of the B-47 aircraft. Side-facing antennas provided considerable pattern distortion. However, a belly installation proved feasible; Figure 3 shows the mounting of the antennas on the B-47 aircraft.

51 The longitudinal slots are horizontally polarized, and the transverse slots give some response to vertically polarized signals. With this set of antennas, left-right indication is necessary to resolve a minor ambiguity. Thus, two side-mounted antennas would be required in addition to the four belly-mounted antennas.

52 Since the slot antennas for sub-band 2 are electrically small, the Q of the antenna circuit is high. Coupling high-Q antennas directly into 50-ohm coaxial systems results in high mismatch losses. Thus, a double-tuned matching circuit was designed to improve antenna performance by about 6 db. A

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matter of practical importance, however, is whether a pair of antennas with matching networks can be phase-tracked to the accuracy required for sub-band 2.

53 It was calculated that two matching networks that differ from each other by as much as 2 percent in a capacitor or an indicator (a shift of 1 percent in resonant frequency) would produce a possible phase mistracking of 1 degree.

54 A pair of antennas and matching networks were built so that their phase tracking could be measured to verify that the assumed variation is a reasonable one for construction of a complete network. Measurement of phase tracking was performed using a hybrid to give sum and difference of the antenna outputs (essentially to change phase information to amplitude, which is more easily measured). Reflections from the roof on which the pair of antennas was mounted in a 15-foot-square ground plane caused inaccuracies in the measurement. Local television and f-m stations caused interference, though a selective receiver was used to reduce this error. Although these conditions reduced the accuracy of measurement, the results indicated that the tolerances calculated to be permissible are ones that can probably be maintained in constructing the networks.

55 Phase-tracking measurements have been made on each of the two pairs of scaled slot antennas for sub-band 2, mounted in the belly of the 1/8-scale model of a B-47. The purpose of these measurements is to determine the probable error contributions of the antennas to the over-all direction analysis. The two antennas were connected to each end of the slotted line. The measurements were performed by probing the SWR pattern on the slotted line.

56 An additional refinement was the use of a super-heterodyne technique. An appropriate local-oscillator signal was fed directly to the probe carriage of the slotted

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line to beat with the energy probed from the standing wave on the slotted line. The resultant i-f signal was amplified in a conventional 30-Mc i-f amplifier, demodulated, and sensed by a conventional SWR indicator. Thus, the sensitivity of the phase-tracking measurement apparatus was greatly enhanced, and the narrow bandwidth of the i-f amplifier (compared with the bandwidth of a single-tuned probe with an audio output) rejected extraneous signals that had interfered with previous measurements.

57 Phase measurements were made on the two pairs of slots farthest aft at scale frequencies of 480, 600, 750, and 1000 Mc, corresponding to full-scale frequencies of 60, 75, 94, and 125 Mc, and at depression angles of 0, 15, and 30 degrees. The data at 15 and 30 degrees were corrected for depression-angle error, inherent in this type of measurement, to determine the antenna effects alone.

58 The measurements were made by feeding the antennas into each end of a slotted line, and then noting the null shift as a function of model position. The results were tabulated as bearing error at various bearings.

59 The points of maximum error correspond to points of poor amplitude tracking of the antenna gain patterns. At these points, the direction analyzer would produce no responses, because the minimum-amplitude selector effectively requires amplitude tracking of about 2 db. By eliminating the points of large error that are rejected by the minimum-amplitude selector, we find rms bearing errors vs frequency of about 6 to 8 degrees, with maximum excursions to 12 degrees (Figure 4) and rms errors vs true bearing angle of about the same order but with maximum excursions to 13.5 degrees (Figure 5).

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(2) SUB-BAND 5

60 Side-facing spiral antennas are used to cover the 480 to 1000 Mc sub-band. These antennas are 6-3/4-inch-diameter, equal strip width, equal spacing spirals printed on 1/8-inch Synthene G-10 and are backed by 2-1/2-inch-deep truncated conical cavities. Each antenna of the pair (the spirals are printed in pairs) had satisfactory polarization ellipticities, radiation patterns, and impedances over the sub-band (references 3, 4, and 5).

61 Phase-tracking measurements were made on these antennas at three frequencies in the 480 to 1000 Mc sub-band. The measurements were made with a slotted line and probe used as a null detector to determine the error in direction finding resulting from antenna structures alone. The antennas were mounted in an 8 foot by 8 foot ground plane. This represents a more perturbed wave-front situation than would be encountered in a fuselage installation, because reradiation from the edges of the ground screen produces ripples in the radiation patterns.

62 Figure 6 shows a mock-up of a section of the B-47 aircraft, showing the spiral antennas mounted on the side of the mock-up. The rms bearing error vs frequency is shown in Figure 7, and the rms bearing error vs true bearing is shown in Figure 8. Measurements of direction-finding accuracy of the system are described in paragraphs 367 and 368.

(3) MEASUREMENT PROBLEMS

63 Considerable interference from roof reflections occurred when both receiving and transmitting antennas were located on the roof of the same building. To reduce these reflections, an improved antenna range (Figure 9) was devised. The transmission path is between the roofs of two buildings separated by about 75 feet. Although this range was not completely "clean," consistent reproducible data were taken.

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64 Gain measurements were made on the low-frequency antenna by means of the three-antenna method, which makes use of the relationship

$$G_t G_r = \frac{P_r}{P_t} \left(\frac{4\pi R}{\lambda} \right)^2 \quad (1)$$

where

G_t = gain of transmitting antenna over that of an isotropic radiator

G_r = gain of receiving antenna over that of an isotropic radiator

P_r = effective power received

P_t = power radiated

R = separation of the antennas

λ = wavelength

Three antennas are operated two at a time in a transmitting-receiving relationship, with means of determining the radiated and received powers. From the measurements of all three possible combinations, a set of three simultaneous equations can be established, based upon equation 1. In these equations, the antenna gains are unknowns, and the other data are established by the measurements taken. Thus, the gain of an antenna under test can be determined through the use of two additional, uncalibrated antennas.

65 Relative-phase measurements at lower frequencies were hampered by the lack of good slotted-line equipment. We chose instead to use hybrid structures to convert phase information to amplitude information, which is more easily measured. Broad-band hybrid structures can be used as sum and difference networks. This measurement procedure provided a satisfactory method of making low-frequency relative-phase measurements.

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4. TELEVISION-INTERFERENCE SUPPRESSORa. INTERFERING-SIGNAL STUDY

66 The frequency range of the AN/DLD-2(XA-1) encompasses the various commercial television bands allocated throughout the world. Television signals include pulses having a high recurrence frequency. Many television stations have effective radiated powers that are comparable with those from low-power radars operating in the same frequency range. Because of these similarities between television and radar signals, circumstances may easily be encountered under which television may provide serious competition with radar throughout a system such as the AN/DLD-2(XA-1). This would be particularly true near large cities capable of supporting several television stations and having substantial radar activity as well.

67 Therefore, it is necessary that the system design of the AN/DLD-2(XA-1) take into account the possibility that television signals may tend to exclude (or reduce seriously) the acquisition of data on radars.

68 Television may cause the following adverse effects within the system:

1. It may tie up memory space in the computer.
2. Because of the high duty cycle of synchronizing pulses and the finite recovery time of the system, it may shut the system down for a good portion of the time.
3. Because direction finding is performed by circuits common to all frequency bands, the presence of complex video signals between synchronizing pulses may cause errors in the direction-finding determination for signals far removed in frequency from the television signals.

69 Therefore, we believe that a satisfactory solution must provide for the elimination, or at least reduction, of the effect of interfering signals at as early a place in the

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system as possible before complete blocking can be produced. As a basis for such a solution, we have investigated two facets of the problem. One facet involved consideration of the various standards used for commercial television, the broadcast frequencies used, and the effective radiated powers produced. The other facet is the make-up of a television signal at the detector output of a crystal-video receiver. Fortunately, all commercial television standards are sufficiently similar that a study of a typical U. S. signal also provides a good general idea of foreign signals.

70 We have found that the frequency limits of present foreign television broadcasting are 41 to 223 Mc, a range that affects the lowest three sub-bands of the AN/DLD-2(XA-1). Most of the stations have effective radiated powers, for the visual component of their signals, of about 100 kw. Although in terms of the present problem the various television standards are similar, there are important differences in synchronizing rates.

71 To study a typical television signal, we set up the receiving system shown in Figure 10. The wide-band amplifier is a Spencer-Kennedy Laboratories type 212C TV Chain Amplifier, covering 50 to 220 Mc. The filter is a breadboard model of the channel 13 filter (75.60 to 81.65 Mc) of the AN/DLD-2(XA-1). This frequency channel corresponds closely to U. S. television channel 5 (76 to 82 Mc) on which WABD-TV broadcasts in New York City; the filter was retuned accordingly. This receiving system is essentially the same as one channel of the frequency analyzer of the AN/DLD-2(XA-1).

72 Our study indicated that most radar signals are considerably more powerful than television signals; hence our receiving range for the radar signals is appreciably greater than for television. At most ranges, television signals should not cause interference. When television signals

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interfere, we decided to introduce attenuation into the antenna lines. Thus, interference is eliminated, and stronger radar signals can be received. Recognition of television signals will depend upon their closely controlled synchronizing-signal rates.

b. PRINCIPLE OF OPERATION OF SUPPRESSOR

73 Figure 11 is a simplified block diagram of the television-interference suppressor. The output pulses from the frequency receiver are standardized pulses that are combined in an OR gate and coupled to the block labeled "synchronizing-pulse recognition circuits." The standardized pulses are applied in parallel to three tuned amplifiers, which are tuned to 10,125 cps, 15,750 cps, and 20,475 cps. These frequencies are exactly or very nearly those used for horizontal synchronization in the various current standards of television broadcasting. When a television signal is received, one of the tuned amplifiers produces an output that triggers the succeeding circuits.

74 The attenuator control circuits operate in the following manner. The output from the above-mentioned tuned amplifier triggers a blanking circuit and a counter, or stepping circuit. The blanking circuit gates off the input from the frequency analyzer for the time required to actuate the electro-mechanical attenuator switches. The stepping circuit controls the relays so that successive steps of attenuation are introduced until the television interference is suppressed. The first step of the stepping circuit initiates a thermal timer, which resets the interfering-signal suppressor after an interval of slightly over 1 minute. If television interference is found again, after resetting, the suppressor renews its cycle; the stepping circuit introduces 10 db of attenuation immediately after the antennas. If 10 db is insufficient, the stepping circuit then introduces 20 db and, if necessary, finally disconnects the antennas and shorts the r-f circuitry of the

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sub-band to ground. About 1 minute after the cycle begins, the suppressor is reset once more, and the antennas are connected directly, without any attenuation from the suppressor. Figures 12 and 13 are schematic diagrams of the television-interference suppressor synchronizing-pulse recognition circuits and attenuator control circuits respectively.

c. LIMITATIONS

75 A television-interference suppressor, functioning according to the principles illustrated in Figure 6, operates in only one sub-band. This restriction is quite satisfactory with respect to the frequency analyzer, which is frequency-sensitive by its very nature, so that one sub-band of this analyzer will not be adversely affected by interfering signals in another sub-band. On the other hand, the direction analyzer (in the three sub-bands using distributed r-f amplifiers) will often tend to produce responses in all three sub-bands to a signal lying in only one sub-band. For pulses other than television, no difficulty results from this situation, since the frequency-analyzer output is used to gate off the incorrect bands until the direction finder has made a selection. For television signals this is not the case, since the signal in the correct frequency band is suppressed so that no such gating signal is generated.

76 Fortunately, this situation is ameliorated somewhat by the antennas and the detectors. The antennas require matching sections between their terminals and the transmission lines leading from them. The combinations of antennas and matching sections, and the detectors, cut off at 12 db per octave outside the sub-bands in which they are intended to operate, and thereby reduce interference from signals lying outside the frequency sub-band of interest. If the total cutoff rate of 24 db per octave is insufficient to eliminate most interference (a matter which we must examine

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from a statistical viewpoint), additional filtering must be provided.

d. ELECTRONIC SWITCHING

77 Electromechanical switching of steps of attenuation is slow and bulky. Electronic switching provides a better means of performing this same function. Switching in sub-bands 1, 2, and 3 can be accomplished by applying appropriate bias voltages to the grid lines of the r-f distributed amplifiers used in those sub-bands. In sub-bands 4 and 5, television-interference suppression, if necessary, can be accomplished by means of fast-acting diode attenuators, described in paragraph 108.

5. DIRECTION ANALYZER

a. MATHEMATICAL DESCRIPTION

78 The direction-finding system uses phase-comparison in its determination of angle-of-arrival. The following mathematical exposition defines the requirements imposed upon the instrumentation.

79 Referring to Figure 14, and assuming two free-space omnidirectional antennas (A and B), the following qualitative picture can be described. At some point along the probed line there is a location where the electrical length through the probed line and lead-in to antenna A plus the air path length to the phase front is exactly equal to the electrical length of the probed line and lead-in to antenna B, located at the phase front. Since these electrical lengths are equal distances from the phase front (a locus of equiphase points), signals at this selected probe will also be in phase (or out of phase if one antenna lead-in is reversed). If the signals are out of phase, complete cancellation will produce a null on the line at this point. Location of this null is dependent upon the time-of-arrival difference of the

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signal at the two antennas. Since time-of-arrival is independent of frequency, the theoretical system is also independent of frequency.

80 Mathematically, it can be stated that the voltages at antennas A and B are given by

$$\begin{aligned} E_A &= |V|e^{-j\beta d \sin \theta} \\ E_B &= |V|e^{-j0} \end{aligned} \quad (2)$$

where

θ = angle-of-arrival

$\beta = 2\pi/\lambda$

d = spacing between the two antennas

λ = free-space wavelength of the received signal

The voltage at some point X on the transmission line joining the two antennas (the voltage magnitude $|E_n|$) is given by

$$|E_n| = 2|V| \sin \left[\beta/2(2X - d + d \sin \theta) \right] \quad (3)$$

where it is assumed that the transmission-line propagation velocity is the same as that for free space, and the transmission-line electrical length is d .

81 Equation 3 is derived by taking the difference output at tap point X. We are concerned with the null; hence,

$$2X - d + d \sin \theta = 0 \quad (4)$$

defines the relationship between position along the line and angle-of-arrival. Note that the measurement is not frequency-sensitive. If $d < \lambda/2$, the only ambiguity present in the

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measurement is the mirror ambiguity, which is defined by $\sin \theta = \sin (\pi - \theta)$.

82 Since the probing of the line is not continuous, it will be instructive to consider the mechanism of selecting the most probable position of the null. Let

- θ_a = angle-of-arrival corresponding to a true null at station a
- θ_b = angle-of-arrival corresponding to a true null at station b
- $\theta_a - \theta_b = 24$ degrees
- θ_1 = angle-of-arrival which lies between θ_a and θ_b such that either $\theta_1 - \theta_a$ or $\theta_1 - \theta_b = 6$ degrees after manipulation, to a first-order approximation

Then,

$$\frac{E_a}{E_b} \approx \frac{\sin 1/2 (\theta_1 - \theta_a) \cos 1/2 (\theta_1 + \theta_a)}{\sin 1/2 (\theta_1 - \theta_b) \cos 1/2 (\theta_1 + \theta_b)} \quad (5)$$

83 Thus, to this order of approximation, the minimum selection ratio is independent of frequency. It is sensitive to the magnitude of the angle-of-arrival. However, if we restrict the region of coverage to -60 degrees $\leq \theta \leq 60$ degrees, an average selection ratio of about 6 db is established, corresponding to a channel width of ± 6 degrees.

84 Further mathematical analysis of the direction-finding system was made to establish requirements on antennas and r-f components (see references 2 and 3). It would be well here to summarize the findings as follows:

1. A left-right indication means is necessary to resolve the mirror ambiguity.
2. Coupling between antennas in a pair must be less than 30 db.

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3. Paired r-f components must track each other in phase to ± 3 degrees--that is, antenna pairs must phase-track to this order, and all other components, such as r-f amplifiers, must phase-track each other to the same order so that direction-finder system error is less than 6 degrees.
4. Paired r-f components must track each other in impedance (SWR) to less than 8 percent.
5. The over-all transmission gain associated with one antenna and its associated r-f circuitry must track the gain of the other antenna and its associated r-f circuitry to better than 2 db.

b. SYSTEM DESCRIPTION

85 Figure 15 is a block diagram of the direction analyzer. In the lowest-frequency sub-band (30 to 60 Mc), the direction analysis is limited to determining whether an intercepted signal arrives from the left or right of the aircraft. This limitation is imposed by the antennas; the requirements that antennas have suitable patterns for accurate direction finding in this frequency region and that they be aerodynamically acceptable are mutually exclusive.

86 Since the antenna patterns in the higher bands are approximately cosine patterns with reduced back lobes, only hemisphere coverage is available with a single antenna pair. In the system, four antennas are used to obtain the coverage and to resolve the left-right ambiguity inherent in the system.

87 The outputs from the four antennas are combined in pairs (by means of hybrids) to produce two signals, as in the basic two-antenna system--that is, the two forward-antenna outputs are combined and the two aft-antenna outputs are combined. These two signals are applied (in phase opposition) to opposite ends of a transmission line and produce a standing wave on the line, which is probed at discrete intervals. The approximate location of the null in the standing wave is

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obtained by comparing the outputs of the detector probes. The probe giving the smallest output is chosen as the most probable position of the null and determines the quantized angle-of-arrival. The additional circuitry involved includes video preamplifiers, logarithmic video amplifiers, a minimum amplitude selector, and an ambel. The left-right indicator, which operates from signals obtained ahead of the combining of forward and aft antenna outputs, serves to eliminate the mirror ambiguity.

88 The left-right ambiguity-elimination circuitry is common to all sub-bands because the system operates on a pulse-by-pulse basis. The pulse-width-analyzer signal is obtained from the left-right indicator. The pads shown in Figure 15 serve to reduce impedance mistracking of the signal lines to an acceptably low value. Part of the signal energy that feeds the probed line is used to drive the frequency analyzer.

89 There are several limitations to the above-described system. In each of sub-bands 3, 4, and 5, back lobes of the spiral antennas are undesirably large. The phase of signals received on the back lobes of antennas on one side of the aircraft may be entirely different from those received on the main lobes of antennas on the other side of the aircraft.

90 The back-lobe level is sufficiently low so that right-left indication can still be obtained. Direction determination, being dependent upon phase, is much more critical. The back-lobe level is sufficiently high that the combination of these signals (of unpredictable phase) with the main-lobe signals of correct phase can cause an error in the direction determination. To avoid this error, the left-right indicator will be used to control r-f gates so that the appropriate antenna pair will be connected to

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the probed line. In sub-band 3, the r-f gates would consist of simple, low-gain r-f amplifiers. In sub-bands 4 and 5, the r-f gating function would be supplied by fast-acting diode attenuators (see paragraph 108).

91 In sub-band 2, the antennas are not bipolarized. Thus, combining the linearly polarized antennas would result in a large direction error, particularly during measurement of cross-polarized signals. This difficulty can be resolved by means of an additional polarization indicator (identical with the left-right indicator). This sensor will select that antenna pair that delivers the strongest signal and gate the appropriate antenna pair to the probed line. The gates would consist of simple low-gain r-f amplifiers.

c. R-F INSTRUMENTATION

(1) DETECTOR MOUNTS

92 Since each sub-band must be operated over a 2:1 frequency range, it was necessary to develop broad-band detector mounts, which are double-tuned networks containing transformation of the crystal-detector impedance (about 3.5 k to 50 ohms). These mounts exhibit a 3-db peak-to-valley Tchebycheff response. The detector mounts for sub-bands 2 through 5 use autotransformers to obtain the impedance transformation. The detector mount for sub-band 1 uses an unsymmetrical network configuration to obtain the impedance transformation. Figures 16 through 20 show the sensitivities of these mounts relative to a tuned detector mount.

(2) HYBRID RINGS

93 Hybrid rings are widely used in the direction analyzer, both to combine and to split signals, where isolation between joined sources or divided loads is required. The hybrid ring is a four-port device, consisting of four quarter-wave transformers, one of which also provides a broad-band

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phase reversal. Hybrids for bands 1 and 2 are constructed with lumped elements. The quarter-wave lines are symmetrical lattice configurations. Figure 21 is a schematic diagram of the lumped-constant hybrid ring. Figures 22 and 23 show isolation and power split of the hybrids for sub-bands 1 and 2, respectively.

94 The coaxial hybrid ring uses a two-wire line for the broad-band phase-reversing section. The capacitive discontinuities in this crossover arm are compensated for in the design. Figure 24 is a photograph and Figure 25 is a schematic diagram of the coaxial hybrid ring. Figures 26 and 27 show the isolation and power-split characteristics of the hybrid for sub-band 4. Figures 28 and 29 show the isolation and power-split characteristics of the hybrid for sub-band 5.

(3) DIRECTIONAL COUPLERS

95 As noted in paragraphs 98 to 103, singly probed lines have distinct advantages over multi-probed lines in the d-f instrumentation. One of the techniques of singly probing utilizes directional couplers to provide the power division and required isolation. For the low-frequency sub-bands, lumped-constant configurations were investigated. These configurations consist of a pair of two-section lumped-constant transmission lines that are both inductively and capacitively coupled. Figure 30 is a schematic diagram of the basic lumped-constant directional coupler, which provides a compact, stable structure and is capable of high directivity and broad-band coupling. Figures 31 through 33 show the isolation and transmission of 3-db, 4.8-db, and 6-db directional couplers, respectively, for sub-band 2.

96 For the higher-frequency bands, the directional couplers were built using slab-transmission-line techniques. Two transmission lines, quarter-wave long at midband, are placed in controlled proximity between a pair of ground

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planes. The impedance of each line and the coupling impedance determine the properties of the directional couplers. Both lumped-constant and distributed-constant directional couplers are well matched over the octave frequency band. Figures 34 through 36 show the isolation and transmission of 3-db, 4.8-db, and 7.5-db directional couplers, respectively, for sub-band 5.

(4) MULTI-PROBED LINES

97 As noted in paragraph 87, the multi-probed line is used to establish the position of the null signal. This position is then related to direction-of-arrival of the incoming signal. The multi-probed lines consist of coaxial lines, tapped at discrete intervals. These taps couple signal energy through minimum-loss pads from the line to the broad-band detector mounts. The minimum-loss pads provide the required decoupling of the detector mount to the line (about 25 db), and the impedance match between the line and the detector mount. Probed-line assemblies were built for sub-bands 2 and 5 for the feasibility demonstration. Figures 37 and 38 show the amplitude and frequency characteristics of these assemblies.

(5) SINGLY PROBED LINES

98 Multi-probed lines must have their probes highly decoupled from the lines to prevent distortion of the standing-wave pattern on the line. Distortion of the standing-wave pattern can result in considerable measurement error. The consequent degradation of system sensitivity is, of course, very serious. It is very desirable to reduce the signal loss introduced in the sensing of the standing-wave pattern.

99 Two methods have been suggested for sensing standing waves, without attendant large decoupling. In both methods, the power from the r-f amplifiers is fed through dividing devices to a multiplicity (eight, for example) of

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transmission lines. The dividers isolate the probed lines so that the probes cannot interact and thereby distort the standing-wave pattern of the system. The probes are arranged symmetrically about the electrical centerline of the system. In the first method (Figure 39), symmetrical trees of hybrid rings divide the outputs from the amplifiers equally among the eight singly probed lines. In the second method (Figure 40), each line is fed by a pair of directional couplers. The pair of couplers nearest the amplifiers feeds one-eighth of the input power to the probed line and passes seven-eighths along the main feed line. The next pair of couplers feeds one-seventh of the remaining incident power (one-eighth of the total) to the probed line. Similarly, each pair of couplers feeds to the probed line a fraction of the remaining power equivalent to one-eighth of the total input power (amplifier output).

100 In both arrangements shown, the theoretical minimum of attenuation imposed upon the system is that produced by the 8:1 power division--9 db. In practice, the loss may be as much as 12 db; however, even this figure compares favorably with the minimum of 25 db considered appropriate for the multi-tapped lines. The binary-tree method has the advantage that all the power dividers are the same, whereas the directional-coupler method requires seven different types. On the other hand, directional couplers are lighter and more compact than hybrid rings. Furthermore, the binary-tree method is limited to systems in which the number of probed points is a power of two; the directional-coupler method is not as limited, and would be adaptable to other systems based upon this development that might require a number of probes other than a power of two.

101 An experimental array, using only three pairs of couplers, rather than a full complement of eight, was built to test the directional-coupler method. Figure 41 is a

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block diagram of the experimental setup used in sub-band 2. The power match of the directional-coupler pairs is extremely good, as shown in Figure 42. The phase tracking of pairs was deemed acceptable, as shown in Figure 43.

102 In addition, we have built a simplified version of the coupler-fed direction analyzer for sub-band 5. The simplified version again uses three sets of couplers; Figure 44 shows the internal construction of the experimental assembly, and Figure 45 shows this assembly with the cover in place. In operation, dummy loads are connected to the output connectors (see Figure 45). Figures 46, 47, and 48 show the SWR patterns measured on this assembly at three frequencies. The pattern asymmetry at 500 Mc has been traced to the inability of the measuring probe to tune to frequencies as low as 500 Mc.

103 These investigations were curtailed due to lack of time. We consider that the results to date are very encouraging, and that this method of singly probing merits further investigation.

(6) DISTRIBUTED AMPLIFIERS

104 Because of probe-decoupling requirements and the sensitivity reduction inherent in null measurements, broad-band (octave) r-f amplification is necessary to obtain the desired receiver sensitivity. Measurements of gain and phase tracking were made on a pair of commercial-quality Spencer-Kennedy type 212-C distributed amplifiers. In the basic phase-tracking measurement, a signal was split, each half amplified in one of the distributed amplifiers and applied to opposite ends of a slotted line (Bone Mfg. Co.). Location of the null by the tuned probe showed phase tracking. Because of the limited tuning range of the slotted-line equipment, the measurements were made from 140 to 220 Mc.

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105 Figure 49 shows the gain tracking of the two distributed amplifiers vs frequency. These amplifiers begin to saturate at a power input level of 0 dbm. Figure 50 is a plot of phase tracking vs frequency at this power input level. The maximum peak-to-peak error is 6 degrees. It was anticipated that phase-tracking error would occur as a result of saturation in the amplifiers. Figure 51 shows the phase-tracking error in the saturation region and the effects on error as the linear region of operation is extended by means of grid bias. It is seen that distributed amplifiers retain good phase tracking well into the saturation region. In general, these results are very encouraging, and we believe that the application of these amplifiers to the lower-frequency sub-bands (1 to 3) will be straightforward.

(7) TRAVELING-WAVE-TUBE AMPLIFIERS

106 Sub-bands 4 and 5 will require traveling-wave-tube (TWT) amplifiers to provide the required r-f amplification. A pair of tracked TWT's was purchased from Huggins Laboratories for test and for use in the feasibility demonstration.

107 Figure 52 shows the amplitude-tracking measurements made on this pair of TWT's. Phase-tracking error was measured by means of a hybrid-ring power splitter and phase reverser and a slotted line as in the measurement of distributed amplifiers. Figure 53 is a plot of phase-tracking error vs frequency. These data show maximum phase errors of 12 degrees and an rms phase error of about 6 to 8 degrees. A second pair of improved TWT's was purchased. However, time did not permit detailed measurements on the second pair. It appears that acceptable phase tracking of a pair of TWT's can be provided. However, considerable phase error is introduced when the TWT's are operated in the saturation region. Beam-gating techniques to extend the linear region were very unsuccessful because of extreme difficulty in controlling

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the phase. Thus, some form of fast-acting controlled-phase attenuation is required to increase the linear dynamic range of the TWT's.

(8) FAST-ACTING ATTENUATOR

108 We built an experimental two-stage attenuator for sub-band 5. Each stage uses two 1N98 crystal diodes between the center and outer conductors of a coaxial line; the mounting of the diodes is such as to minimize their reactance. The two stages are spaced one-quarter wavelength apart at the center frequency of the sub-band. Figure 54 shows the experimental fast-acting attenuator, disassembled. Figures 55 and 56 show the impedance and attenuation characteristics as a function of control current and frequency. A five-section attenuator will have a maximum attenuation of greater than 30 db and a minimum attenuation of only 1.5 db. We believe that this device will substantially solve the dynamic-range problems associated with TWT's in sub-bands 4 and 5.

d. VIDEO INSTRUMENTATION

(1) LOG VIDEO AMPLIFIERS

109 Log video amplifiers are used in the direction analyzer, because relative amplitudes must be preserved over a very wide dynamic range. The amplifier uses common-emitter stages with gain-switching networks in the emitter circuits, as required, to provide the approximately logarithmic characteristic of the device.

110 Since a common set of log video amplifiers is used for all four sub-bands, there are four preamplifiers corresponding to each direction channel. Each preamplifier has an individual gain control to compensate for detector-crystal variations. The common post-amplifier consists of a linear stage, five gain-switched stages (to produce the logarithmic characteristic), and an emitter follower to provide a low-

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impedance output. A gain control is incorporated in the linear stage to compensate for transistor variations.

111 Figures 57 is a schematic diagram of the preamplifier. Figure 58 is a schematic diagram of typical gain-switched stages. Gain switching is produced by diodes in the emitter circuits. Under no-signal conditions, each diode is forward-biased, producing a small amount of emitter degeneration. A small signal does not cause sufficient flow of emitter current to overcome the bias; the small-signal gain is, therefore, governed by the small amount of emitter degeneration produced by the conducting diode. A large pulse causes the diode to be cut off; this increases the emitter degeneration so that the stage gain is about unity.

112 Recovery-time reduction is obtained by means of one short-time-constant coupling circuit between the second and third gain-switched stages and a negative-feedback connection from the last gain-switched-stage collector to the third gain-switched-stage base. Since the feedback path includes a diode, the feedback is effective only for the overshoot immediately after the pulse, and hence does not affect the desired pulse itself. Figure 59 is a schematic diagram of the logarithmic amplifier.

113 Recovery time was measured using two pulsed r-f signal generators. One signal generator is modulated by a 100- μ sec pulse at the maximum expected r-f signal level (-10 dbm). The other signal generator is modulated by a 10- μ sec pulse, whose delay with respect to the first pulse is variable, at the minimum expected r-f signal level (-50 dbm). The low-level 10- μ sec pulse is moved toward the high-level pulse until the amplifier low-level output begins to lose amplitude. This occurs about 20 μ sec after the end of the large-signal pulse.

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(2) MINIMUM-AMPLITUDE SELECTOR

114 The minimum-amplitude selector (MAS) serves to determine the amplifier that has the smallest signal at its log-video-amplifier output, thereby establishing the most probable position of the null on the probed line.

115 The inputs (eight) are applied to an OR gate, which passes the largest signal. The largest signal is then fed back to a subtracting amplifier in each channel input of the MAS, when the largest input is subtracted from all others (including itself). The outputs from the subtracting amplifiers are a group of pulses whose amplitude rank is the inverse of the input group: the smallest has been converted to the largest; the next-to-smallest to the next-to-largest, etc. The new set of pulses is applied to a peak-amplitude selector, which produces a standardized output from a blocking oscillator in the channel corresponding to the largest input.

116 The largest pulse that is applied to the peak-amplitude selector is passed by an OR gate; a bias applied to the gate diodes causes the largest pulse to be clipped by a small fixed amount (independent of the amplitude of the pulse). The amplifier has a gain of exactly -1. The clipped, inverted pulse is applied to a resistive mixing circuit in each channel. The other input to each mixer is the channel input pulse, and the mixer output is the difference between these inputs. In the channel whose individual input is the largest pulse, the mixer output is a pulse of the same polarity as the input, with an amplitude equal to the amount clipped at the input to the inverting amplifier. In all other channels, the polarity of the mixer output is the opposite of the polarity of the individual input, because the clipped, inverted pulse corresponds to the largest of the individual inputs.

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The mixer outputs are amplified, but only the output that is generated in the channel having the largest input (the smallest input to the MAS) has the proper polarity to trigger the blocking oscillator. If two inputs to the peak-amplitude selector are very nearly the same in amplitude, within the clipping level associated with the inverting amplifier, both may produce outputs. Thus, the clipping level determines the minimum-amplitude selection ratio.

118

Figure 60 is a block diagram of the MAS. The subtracting amplifier is labeled the active difference circuit, to emphasize its distinction from the resistive mixer or passive difference circuit (see Figure 61). The buses that connect the common circuits with the circuits repeated in several channels are labeled with Greek letters, to distinguish them from one another. The α bus carries the largest input signal to a level shifter, and the β bus returns the level-shifted signal to the active difference circuits, where the amplitude rank of the inputs is inverted. The γ bus carries the clipped signal corresponding to the largest input to the peak-amplitude-selector section; the γ bus signal is applied to the -1 amplifier, and the output from the -1 amplifier is returned to the passive difference circuits via the δ bus.

119

The α bus not only carries the largest of the inputs to the MAS, but also responds to the input (above a predetermined threshold) that arrives first. Note that relative delay through all direction channels is not constant, but varies as a function of level, pulse rise time, etc. The first signal is delayed by 0.9 μ sec (a time shown experimentally to be sufficient to cover differences in pulse delay time), after which it triggers a blocking oscillator. This threshold device establishes the sensitivity of the direction analyzer. The pulse generated by this blocking oscillator is applied

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via the ϵ bus to the output blocking oscillators of the individual channels. Thus, clocking is provided by the ϵ bus so that no error is introduced as a result of variations in pulse delay.

120

Figure 62 is a simplified schematic diagram of the MAS. Note that passive mixing introduces pulse amplitude loss. This loss can be tolerated, because amplification in each channel precedes the output blocking oscillator. However, loss cannot be tolerated in a cascade because of the necessity of preserving levels to a high order of accuracy. Thus, an active subtracting circuit is used in the first function (despite cost) and a passive subtracting circuit is used in the second function (where pulse polarity provides the decision).

(3) LEFT-RIGHT INDICATOR

121

The left-right indicator is used to eliminate the problems associated with limited antenna coverage and mirror ambiguity, and to provide the signal drive for the pulse-width analyzer. The basic mechanism of indication uses log video amplifiers and a simple peak selector.

122

The left-right indicator is required to accept signals with a dynamic range of 60 db. The indicator will be common to five sub-bands; hence, preamplifiers will be required to feed the common indicator. Figure 63 is a schematic diagram of the preamplifier for the indicator and the driver for the pulse-width analyzer. The preamplifier differs little from the direction-channel preamplifier.

123

The log video amplifier in the indicator differs from that in the direction channel in that the first stage is also a gain-switched stage (see Figure 64). This change was required to handle the increased dynamic range. The peak selector of the left-right indicator uses cascaded

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differential amplifiers. If unequal inputs of the same polarity are applied, the outputs are equal but of opposite polarity. Thus, if we follow the differential amplifier with a circuit that discriminates between pulse polarities (blocking oscillator), suitable selection can be obtained. Figure 65 is a schematic diagram of the peak-amplitude selector.

124 The imbalances in response time to fast waveforms that occur in differential amplifiers are used to supply a portion of the pulse-width driver circuit. These imbalances result in pips on the leading and trailing edges of the output pulses. A delay line is placed in one input to a differential amplifier, and the pulse input from the preamplifier (Figure 63) is applied to both inputs of the amplifier--directly to one, and via the delay line to the other. The outputs of the amplifier, then, have large pips at the leading and trailing edges of the pulses. The delay mechanism in one input essentially determines the slope of the pulse rise and fall times; when the magnitude of the slopes exceeds a predetermined value, the pips are generated. Figure 66 is a schematic diagram of the driver for the pulse-width analyzer.

(4) AMBEL AND MATRIX

125 Figure 67 is a schematic diagram of the direction ambel and coding matrix; the latter converts the direction-channel information to binary form. The clock drive (from the master clock of the receiving system) is generated in response to the operation of the coincidence gate when that gate receives time-coincident signals from the direction and frequency analyzers. The clock drive is regenerated in a blocking oscillator, which releases a high-current clamp on the coding-matrix output buses.

126 The minimum-amplitude selector chooses the smallest signal. When nulls fall midway between two probes, however,

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equal-amplitude signals are provided, both of which are passed by the minimum-amplitude selector. To handle this situation, an ambel is included.

127 The ambel passes a signal in a single direction channel. When signals are present in an adjacent pair of channels, the ambel inserts an interpolate channel between the two real direction channels. When signals are present in three direction channels, the ambel rejects them all. When signals are present in the fore or aft channels, the signals are rejected.

128 The operation of the ambel depends upon an inhibit gate, which operates as follows. For positive pulse input, in the absence of an inhibiting input, the transistor is biased off and its collector resistance is much higher than the external resistance. Thus, if an input is applied, it is passed to the output with little pulse attenuation. When an inhibiting input is applied, the transistor is switched on and its collector resistance drops to a much lower value than the external resistance. If an input is applied at this time, it is greatly attenuated.

129 There are six inhibit gates in the ambel, corresponding to the six acceptable direction-channel outputs from the MAS. The other two direction channels are associated with the 60-degree sectors directly fore and aft of the aircraft. If a single signal is applied to the ambel by the MAS, it will pass through its inhibit gate and be encoded in binary form by the matrix; if it corresponds to an unacceptable channel, it is not passed to the matrix, and is thus rejected by the system.

130 If two adjacent inputs, one of which corresponds to an unacceptable channel, are applied, the unacceptable input inhibits the acceptable one. Thus, direction information from the edges of the "unwanted" sectors is also rejected

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(a guard function). If two adjacent acceptable inputs are applied, however, the lower-numbered input inhibits the higher-numbered input, and the lower-numbered input is passed to the matrix for encoding. In addition, adjacent acceptable inputs will excite both OR gates connected to the inputs; the lower-numbered output will excite a third OR gate. The three OR gates drive an AND gate whose output, via an emitter follower, is the interpolate pulse. The interpolate pulse indicates that the direction of arrival is halfway between the two adjacent direction channels. Thus, the ambel increases the resolution of the direction analyzer.

131 If three adjacent signals are applied to the ambel, they mutually inhibit one another, and no coded signal is produced by the ambel. The two input OR gates are excited, but the output OR gate is not; thus, no interpolate signal is generated. The system does not admit three adjacent-channel signals as valid.

132 The interpolate and binary outputs trigger blocking oscillators, which transmit the direction information to the computer.

6. FREQUENCY ANALYZER

a. GENERAL

133 Figure 68 is a block diagram of the frequency analyzer. The analyzer uses 45 filters and associated crystal-video receivers to channelize the 30 to 1000 Mc range covered by the AN/DLD-2(XA-1). The video amplifiers drive blocking oscillators to produce standard output pulses. The output pulses are applied to a logic circuit (ambel), which eliminates ambiguous responses and recognizes the blocking-oscillator pulse pairs that are produced by received signals in filter crossover regions. In this case, the ambel generates an interpolate pulse. Thus, virtual channels,

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lying between the real channels created by the filters, are produced, so that the 30 to 1000 Mc range is effectively divided into 89 channels. Data pulses from the ambel representing these channels are passed to the computer. A matrix encodes the frequency data pulses into binary form.

134 The outputs from each group of crystal-video receivers, belonging to a sub-band, are OR-gated, as described in reference 4; the output from any one OR gate switches off the r-f inputs to the other sub-bands and thus reduces interference with direction analysis in its own sub-band. In the lowest three sub-bands, the outputs from the OR gates also control the television-interference suppressors. Although it is believed that there is no necessity for such suppressors in the two highest-frequency sub-bands, there is no reason why these suppressors cannot be added, if required. The output from the OR gate of sub-band 1 also controls the generation of an interpolate pulse to validate pure left-right data that are generated in that sub-band. Finally, signals from the five sub-band OR gates are combined in a sixth OR gate and used to drive a central clock, which is described in paragraphs 176 through 178.

135 In each of the sub-bands, except sub-band 5, the connection to the OR gate from the highest-frequency channel is made via a delay line and an inhibit gate. Without these provisions, a signal near the crossover frequency of two adjacent sub-bands might produce outputs from the OR gates of both sub-bands and thereby prevent any direction analysis being made of that signal. Double-gating is prevented by the inhibit gates, which permit the upper-frequency sub-band to capture a signal that has a frequency near the crossover of the two sub-bands. The delay lines ensure that this capture can occur though the speed of response of the higher-frequency channel involved may be slightly less than that of the lower-frequency channel.

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136 In the complete analyzer, there are 45 filters, 45 detectors, 45 video amplifiers, an ambel, a matrix, and a set of output blocking oscillators. The filters channelize the 30 to 1000 Mc range into 45 channels, each of which is 7.7 percent wide. The ambel develops an interpolate channel between centers of adjacent filter channels to provide an effective 89-channel selectivity in the 30 to 1000 Mc range.

137 Filters will be connected alternately to an input from the forward pair of antennas of the direction analyzer. The intervening filters will be connected to an input from the aft pair of antennas of the direction analyzer. The separation of the inputs to adjacent filters reduces interaction at the band edges. This simplifies filter design and improves filter operation.

138 The output from each filter is detected and amplified in a video amplifier. This amplified signal is used to trigger a standardized output pulse generator in the video amplifier. An OR circuit in the output from each receiver transmits a gating-off signal to the direction analyzer.

139 If the filters had rectangular pass bands, only one channel would respond to a strong signal. With practicable pass bands, strong signals give responses in one, two, or at most three channels. An ambel is used to select the most probable frequency channel when more than one channel responds. Figure 69 shows the regions of multiple signal responses. If the frequency and amplitude are such that the signal falls in the region marked M, only a single video amplifier will be triggered. For this condition, no ambel action is required, and any signal that falls in this region is called a channel 8 response. If the signal falls in the frequency-amplitude region marked N, two video amplifiers will be triggered. The ambel designates any signal in this region as a response in channel 9. If the signal falls in the region marked P, three

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video amplifiers are triggered. The ambel designates any signal in this region as a response in channel 8. It is expected that no signals that are more than 40 db over the triggering level will appear at the detectors. There should, then, be no signals in region Q. If there were responses in four adjacent channels or two non-adjacent channels, because of noise or more than one signal in exact time coincidence, the ambel would class these responses as unsatisfactory situations and would reject all information on this intercept.

140 The matrix converts the parallel form of channel information to seven binary digits. Output blocking oscillators furnish these binary digits as standardized low-impedance pulses to the computer. To furnish this information in time coincidence with the direction information, the output blocking oscillators are gated from the central clock.

b. FILTERS AND DETECTORS

141 The basic filter configuration has been designed to meet the following requirements:

1. The input impedance outside the pass band will be high to facilitate the parallel connection of a number of filters. (This is conveniently achieved by means of a series-resonant input circuit. Another advantage of the series-resonant input circuit is that it is easily modified to give good pass-band and crossover response when several adjacent band filters are connected in parallel.)
2. The input impedance in the pass band will match the input line impedance of 40 to 50 ohms.
3. The output impedance in the pass band will match the impedance of the crystal that is included as an integral part of the filter. The resistive part of the impedance of the crystal, when forward-biased with about 15 μ a to improve the detection efficiency, has been measured at about 3500 ohms in the 30 to 120 Mc sub-bands.
4. The output will have a shunt-resonant circuit to provide a d-c return for the detector crystal.

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5. The insertion loss in the pass band of the filter will be small.
6. The insertion loss outside the pass band (off-band rejection) will reach about 38 db at one bandwidth outside the pass band. (If the dynamic range at the output from the r-f amplifier is compressed to 40 db, this off-band rejection prevents response in more than three frequency channels.)
7. The electrical values theoretically required for the components can be attained in practice. In this regard, the types of resonant circuits that are chosen at the input and output, in conjunction with their loading impedances, result in convenient component values. The remainder of the circuitry was chosen to accomplish this purpose as satisfactorily as possible.

142 Proper off-band rejection is obtained with a Butterworth filter using four coupled resonant circuits. These filters can be coupled with mutual-inductive, inductive, capacitive, or direct coupling, or a combination of couplings. For a combination of capacitive and direct couplings, the circuit provides a series-resonant input (for proper input impedance) and a shunt-resonant output (for d-c return to the crystal). The filter design is based upon symmetrical loading. The actual difference in impedance levels at input and output is compensated by an ideal transformer at the output.

143 Lumped-constant filters were designed for sub-bands 1 and 2. Figure 70 shows the responses of the r-f filters for sub-band 1, fed from a coaxial feed line of 40 ohms impedance, and terminated in a 1N21 crystal detector. Figure 71 shows the responses of five filters for sub-band 2 operated in parallel. The solid curves represent the responses of the five filters in parallel. The dashed curves, which have been included for comparison, represent the responses of filters operating without any other filters in parallel

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and without being retuned after the test involving parallel operation of five filters. The effect on any one of the filter pass bands produced by the other four parallel filters is minor. Setting up this test involved an investigation of the best means of tuning a number of paralleled filters. This was directly related to a general study of methods of tuning the filter and their effects on filter response.

144 Two methods for tuning the filter were considered. Both methods begin with all resonators detuned. In the first method, which will be called forward tuning, power is supplied to the filter input in the normal manner and the first (input) resonator is probed with an r-f detector. The four resonators are then tuned in order, starting with the input resonator. The second method of tuning involves probing the fourth (output) resonator with a loosely coupled source of r-f power to the fourth resonator and observing the output from the crystal detector that normally loads the filter. The four resonators are then tuned in reverse order, starting with the output resonator. To detune the resonator succeeding the one that is being tuned, a short-circuit is placed at an appropriate point.

145 The following is a description of the forward-tuning procedure (refer to Figure 72). In this description, the part of the circuit in Figure 72 indicated as "reverse tuning probe" is removed. The forward-tuning probe is a high-impedance device, such as the r-f probe of a vacuum-tube voltmeter.

Step 1.---Short-circuit C_{2p} , and adjust C_1 to minimize probe output. Remove the short-circuit. This action resonates L_1 with the series combination of C_1 and C_{12} , with a small error (0.25 percent), because the short-circuit has placed L_2 in parallel with C_{12} . Since C_{12} and L_2 are large, $\omega L_2 \gg 1/\omega C_{12}$, and the error is of no significance.

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Step 2.--Short-circuit L_3 , and adjust C_{2p} to maximize the probe output. Remove the short-circuit.

Step 3.--Short-circuit L_4 , and adjust C_3 to minimize the probe output. Remove the short-circuit.

Step 4.--Adjust C_4 to maximize the probe output. This procedure would result in a value of C_4 producing a series resonance (looking into C_{34} from L_3) if the conductive loading of L_4 by the 1N21B crystal were removed. The crystal cannot be removed because it places a capacitance across L_3 and is an appreciable part of the total capacitance C_4 shunting L_3 . Hence, the tuning procedure results in C_4 being adjusted erroneously to produce a maximum admittance looking into C_{34} . The error in the setting of C_4 results in an error in the resonant frequency of the output resonator and in an error in the coefficient of coupling (k_{23}) to the preceding resonator (L_3 and C_3). This is a major error.

146

The following is a description of the reverse-tuning procedure. In this procedure, the part of the circuit of Figure 72 indicated as "forward tuning probe" is removed. The output from the source at the input to the filter is reduced to zero. Coupling capacitance is essentially in parallel with C_4 during tuning ($X \gg R_Q$) and is maintained in this position after completion of tuning by closing switch S. Short-circuiting R_Q of the reverse-tuning probe after tuning produces a slight error (about 0.15 percent) in the tuning of the fourth resonator.

Step 1.--Short-circuit L_3 , and adjust C_4 to maximize the video output. Remove the short-circuit.

Step 2.--Short-circuit C_{2p} , and adjust C_3 to minimize the video output. Remove the short-circuit. The tuning procedure produces a resonance between L_3 and a capacitance equal to C_{2s} , C_3 , and C_{34} , all in parallel. The desired resonance is between L_3 and capacitance equal to C_3 and C_{34} in parallel. Because $C_3 \gg C_{2s}$, the error is small, about 0.93 percent.

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Step 3.--Place a short-circuit to ground between L_1 and C_1 , and set C_1 at minimum capacitance. Adjust C_{2p} to maximize the video output, and remove the short-circuit. There is a small error in the setting of C_{2p} because C_1 has been in parallel with C_{12} during tuning. This error is about 0.03 percent.

Step 4.--Adjust C_1 to produce a minimum video output, and close switch S to complete the tuning. The adjustment of C_1 should produce a resonance between L_1 and the series combination of C_{12} and C_1 . The resonance will be produced only if the resistive loading on L_1 represented by R_g is reduced to zero. If not, a major error results, as in step 4 of the forward-tuning procedure.

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In the discussion of forward- and reverse-tuning procedures, numbers have been assigned to percent errors introduced by the first three steps of tuning. These refer to percent errors in both resonant frequency and coefficients of coupling k_{12} or k_{23} , depending upon whether resonators 1 and 2 or resonators 2 and 3 are being tuned. These errors are relatively small (from 0.03 to 0.93 percent) in comparison with that produced in step 4 with the fourth resonator normally loaded. The latter error is about 11 percent in both f_0 and k_{34} for forward tuning. The error is 6.48 percent in k_{12} and 6.12 percent in f_0 for reverse tuning. Since the conductive loading of the fourth resonator is inseparable from part of the tuning reactance (1N21B crystal capacitance), it is impossible to eliminate this error in forward tuning. In reverse tuning, however, it is possible to short-circuit the source impedance loading the last resonator to be tuned (L_1 , C_1 , and C_{12}) with a short-circuit whose inductance is negligible with respect to that of L_1 . Thus, the error can be almost completely eliminated by reverse tuning.

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In practice, it was found possible to tune the filters with the last resonator to be tuned normally loaded,

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provided that (1) the filters were to be operated individually and not in parallel, and (2) the tuning frequency and fixed components of the filter were adjusted to produce the desired bandwidth and band limits. Examination of the condition of high stop-band input impedance required for parallel operation of several filters shows that the impedance depends essentially upon the input series circuit (L_1 and C_1). This circuit is almost shorted to ground by C_{12} when the frequency is far from the resonance of L_1 and C_1 . In reverse tuning, the input series circuit is the last to be tuned, and its resonant frequency is incorrect by almost one filter bandwidth when the resonator is tuned in its normally loaded condition. Under these conditions, the stop-band input impedance is considerably lower than expected on one side of the pass band. Thus, it is to be expected that filters reverse-tuned in this manner will interfere with each other when placed in parallel. This situation is borne out by experiment.

149 Similar results were obtained with forward tuning. The causes of this similarity are not obvious, because it is the output resonator (L_4 and C_4) that is improperly tuned. It was found, however, that when reverse tuning is used and the loading on the input resonator is short-circuited during tuning, there is little difference in the response whether filters are operated individually or in parallel with a group of similar filters (Figure 71).

150 The design of the filters for sub-bands 1, 2, and 3 was similar. In sub-bands 1 and 2, each individual filter configuration is designed for use at the center channel of a group of three channels in the sub-band. The same filter design is then used for all three channels by tuning the filter to the appropriate frequency. Bandwidth variation is the factor limiting the number of channels over which one filter design can be successfully retuned. When a filter

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design is limited to three channels, the bandwidth stays within 10 percent of the center-channel bandwidth. The configuration of the filter for the three lowest-frequency channels of sub-band 3 is the same as that in sub-bands 1 and 2. In the two filter types that will cover the six upper-frequency channels of sub-band 3, it was necessary to modify the output resonant circuit from that used in the three lowest-frequency channels. Figure 73 shows the responses of the r-f filters for sub-band 3.

151 Lead inductance, stray capacitance, minimum capacitance of tuning capacitors, and low unloaded Q make continuation of the lumped-constant filters into sub-band 4 impractical. The best compromise of unloaded Q , small size, and freedom from spurious responses in sub-band 4 proved to be a filter made of coaxial resonators lump-loaded with tunable capacitors and tunable-capacitor inter-resonator couplings. To minimize the size of these high- Q resonators, a 3-db-ripple three-resonator Tchebycheff filter has been used instead of the four-resonator Butterworth filter used at lower frequencies. The capacitively loaded resonators had an unloaded Q of over 500. This is sufficient to build Tchebycheff filters with an insertion loss of less than 1 db. (With lumped-constant elements at lower frequencies, the Butterworth configuration was required to maintain low insertion loss with unloaded Q 's of 100 to 500.)

152 An attempt to use a single resonator over several channels indicated that retuning of the resonators alone was insufficient because the bandwidth variation was too great. Simultaneous variation of the inter-resonator coupling improved this situation greatly. Thus, two cavity lengths (those for channels 64 and 70) were sufficient to cover all nine channels in sub-band 4. Experimental work on these filters required no more than one substitution of any one

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coupling capacitor to achieve the desired couplings. Figure 74 shows the responses of the filters for sub-band 4.

153 Initial work on the filters for sub-band 4 was accomplished with a single filter body consisting of three variable-length coaxial cavities. Each cavity was loaded with a tubular air-dielectric variable capacitor. Couplings between adjacent resonators (and between the input resonator and a slab-type 50-ohm feed line) were accomplished by means of fixed ceramic capacitors.

154 It was found that the unloaded Q of these cavities ranged well over 1000 after the substitution of the air-dielectric variable capacitor for the ceramic type used in sub-band 4. All nine filters of sub-band 5 were breadboarded individually, using the three-cavity structure. Only two different cavity lengths were used for all nine filters; the shorter length covers the five higher frequencies. Figure 75 shows the responses of the nine individual filters. The peak-to-valley ratios are close to theoretical, and the insertion losses are negligible.

155 Some spurious responses were encountered in the stop bands of these units. These responses were traced to series resonances in the coupling capacitors. The spurious responses were suppressed by replacing single coupling capacitors by parallel combinations having the same total capacitance.

156 When breadboard testing of the single filters was completed, two units were designed to embody four filters in one case and five in the other. All of the filters in each case are fed from a common slab-type 50-ohm feed line.

157 Frequency bands of the four-filter unit alternate with those of the five-filter unit. A theoretical analysis indicates that each of a group of filters so connected to a

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feed line will not be materially affected by the aggregate of the impedances representing the remaining filters on the same feed line (the calculated transmission change is about a few tenths of a db). This analysis was verified when the four- and five-filter units were constructed. There was essentially no difference between the pass-band responses of the original breadboard filters and their duplicates operating with other filters from a single feed line.

158 Some spurious responses showed up, however, in the stop bands; in one filter, in particular, stop-band rejection dropped to as low as 20 db at some frequencies. The minimum rejection in this and all other filters was raised to more than 30 db by the insertion of lossy material at appropriate points along the feed line. It appears that some of the spurious responses may be associated with a non-TEM mode of radiation from the slab-type feed line. Figure 76 shows the assembled filter groups for sub-band 5.

c. VIDEO AMPLIFIER

159 There is one video amplifier for each r-f filter and detector. The desired output of the video amplifier is a standardized pulse corresponding to an r-f input greater than some predetermined threshold.

160 The first amplifier developed contained 2N263 transistors and consisted of four common-emitter stages in approximately iterated cascade, each stage using rather heavy shunt feedback (collector to base) to improve the frequency response and stabilization of the amplifier. This amplifier had acceptable rise time (1 μ sec), the gain of each stage was about 20 db, and its over-all gain was sufficient. The recovery time was about 25 μ sec for a 10- μ sec pulse of 1-percent duty factor at 0 dbm.

161 This amplifier failed badly with wider pulses or higher duty factors. The stages blocked and the original

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method of clamping the adjacent collectors together with diodes resulted in excessive pulse stretching of wide pulses, especially with pulses having poor fall time. A new method for handling the required dynamic range became necessary. The most promising method appeared to be compression, as used in the log video amplifier. The requirements for the frequency-receiver video amplifier, however, unlike those for the log video amplifier, do not make it necessary to maintain uniform compression in all stages. The first two stages of the amplifier were designed to switch (by means of a diode in each emitter circuit) from full gain for small signals to low gain for large signals. The small-signal gain of such a compression stage, however, became somewhat less than 20 db because of increased emitter degeneration caused by the diode resistance. The over-all gain of the amplifier became inadequate, and a fifth stage (of the same basic type as the third and fourth) was added.

162 In this type of high-gain RC-coupled amplifier, the low-frequency overshoots, resulting from imperfect low-frequency response of each stage, create considerable difficulties by overdriving the amplifier and causing changes in the d-c operating conditions. For the original amplifier, a short-time-constant coupling circuit was used between the second and third stages. With the addition of the fifth stage, the short-time-constant coupling circuit was placed between the third and fourth stages, because it was found that the overshoot due to the differentiation of the pulse by the short-time-constant coupling circuit was too difficult to handle after more than two amplifications.

163 The low-frequency overshoots prevented placing the short-time-constant coupling circuit still further back (between the fourth and fifth stages). Even with the short-time-constant coupling circuit between the third and fourth stages, and with compression in the first two input stages,

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the low-frequency overshoots were still excessive. All bypass and coupling capacitors were increased to the limit allowed by practical design considerations. This resulted in some, but not sufficient, improvement. Various feedback methods were tried for further improvement but without much success. Finally, compression was also added to the third stage. Although this approach gave the best results to date, additional improvement was required.

164 The short-time-constant coupling circuit caused a negative overshoot at the trailing edge of the pulse. After two amplifications (in the fourth and fifth stages), this overshoot was sufficiently large to cause blocking in these stages and resulted in a considerable increase in the amplifier recovery time. Diode-clipping circuits in the third and fourth stages were helpful in reducing the overshoots in amplitude and the time constant of the recovery exponential. This was accomplished, however, at a cost in gain by a factor of about 2.5.

165 Figure 77 is a schematic diagram of the video amplifier for the frequency analyzer. The low-frequency response of the amplifier has been improved (thereby reducing overshoots) by elimination of the emitter-circuit coupling capacitors in the gain-switched stages. (This elimination is not possible in the log video amplifier of the direction analyzer. With direct coupling between the emitter of the gain-switched transistor and the switching diode, there is d-c interaction that prevents the close control of diode bias current necessary for accurate shaping of the transfer characteristic in the log video amplifier. The bias current in the video amplifier need not be controlled so closely.)

166 As a basis for eliminating the emitter-circuit coupling capacitors, we made a series of measurements of the variations of d-c emitter voltage in individual test stages.

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The test stages were operated without a-c signals, and the d-c effects of changes of transistors and of temperature were measured. These experiments established that the emitter voltage of the type of amplifier stage used here is adequately stabilized by the combination of emitter and collector-to-base feedbacks. Diode characteristics as a function of temperature were also measured.

167 A diode is placed at the input terminal of the amplifier to limit the input signal and reduce overloading on large signals. A frequency-compensating network in the emitter circuit of the fifth stage sharpens the trailing edge of the overshoot and thus improves the recovery time of the amplifier.

168 The tangential sensitivity of the amplifier, observed at the collector of the fifth stage, is -53 dbm at room temperature. The amplifier was fitted with transistors and diodes selected at random for this test, and the detector crystal was tuned. After this test, the threshold sensitivity (signal required to produce an output from the blocking oscillator) was measured. The threshold is set as follows. The threshold control is adjusted to trigger on noise once per second. A signal is introduced until a 50-percent probability of firing occurs. The signal level is increased by 3 db, and the threshold control is readjusted until a 50-percent probability of firing is obtained. When the threshold control is reset to introduce the added 3 db of protection, the false-alarm rate is reduced to appreciably less than one per minute. Under these conditions, the sensitivity is -46.5 dbm for 1 to 100 μ sec pulses.

169 Recovery time was measured by a two-pulse technique, as used with the log video amplifier. For a first pulse, 20 to 100 μ sec wide and at a power level of 0 dbm, the threshold sensitivity to a second pulse, 1 μ sec wide, was 1 db

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down--that is, -45.5 dbm--25 μ sec after the end of the first pulse and 3 db down 20 μ sec after the end of the first pulse. The recovery time after a 10- μ sec-wide pulse was slightly shorter. The amplifier operated satisfactorily over the required temperature range.

d. AMBEL AND MATRIX

170 The ambel has four input buses. The video amplifiers are coupled cyclically to the buses in the order ABCDA, etc. The requirements of the ambel are as follows:

1. When there is a single input, the corresponding output bus shall be energized.
2. When there are inputs on two adjacent channels, the output shall appear on that bus corresponding to the lower input (in the order ABCDA) and, in addition, on an interpolate bus.
3. When there are inputs on three adjacent buses, the output shall appear on the output bus corresponding to the middle input.
4. When there are inputs on all four buses, there shall be no output.

171 We designate the output buses corresponding to A, B, C, and D as A', B', C', and D', respectively, and the interpolate bus as I; the requirements of the ambel can then be summarized in the truth table (see Table II).

172 These requirements can be expressed in Boolean algebraic form without intermediate steps; if intermediate steps are used, the expressions are simplified. In addition, the conversion from algebraic expressions to circuits that perform the functions indicated by the expressions yields an instrument with fewer gates than would be required to perform the ambel function in a parallel, rather than in a cascaded manner. The Boolean equations, including an intermediate step yielding the double-prime notation, are as follows:

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TABLE II
TRUTH TABLE FOR FOUR-BUS AMBEL

Inputs	Outputs
A	A'
B	B'
C	C'
D	D'
AB	$A'I$
BC	$B'I$
CD	$C'I$
DA	$D'I$
AC	None
BD	None
ABC	B
BCD	C
CDA	D
DAB	A
ABCD	None

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First Law:

$$\begin{aligned} A\bar{C} &= A' \\ B\bar{D} &= B' \\ C\bar{A} &= C' \\ D\bar{B} &= D' \end{aligned} \quad (6)$$

Second Law:

$$\begin{aligned} A'\bar{D}' &= A' \\ B'\bar{A}' &= B' \\ C'\bar{B}' &= C' \\ D'\bar{C}' &= D' \end{aligned} \quad (7)$$

Interpolate Law:

$$\begin{aligned} A'B' B'C' C'D' D'A' &= I \\ \text{or } (A' + C')(B' + D') &= I \end{aligned} \quad (8)$$

Figure 78 is a block diagram and Figure 79 is a schematic diagram of the frequency-receiver ambel.

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The ambel uses the inhibit gate and conventional diode gates in its construction. The first set of inhibit gates (nearer the inputs) mutually eliminates non-adjacent signal inputs in accordance with the first law. The second set of inhibit gates eliminates the higher of an adjacent signal pair in accordance with the second law. The interpolate law is given effect by the cascaded OR and AND gates that are tapped off between the first and second laws. The ambel output signals are regenerated by means of clocked blocking oscillators.

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174 Time delays, in passage of pulses through gates, are subject to considerable variation. Depending upon the combinations of applicable ambel laws, coincidence of pulse-rise times may be lost at the inputs to inhibit gates. A short time after the pulse rise, however, all logical functions will have been performed. Thus, the generation of outputs is delayed by the delay line and clock blocking oscillator. This technique prevents erroneous output information. The ambel operated satisfactorily over a -60 to +90 C temperature range.

175 The frequency-coding matrix converts the frequency-channel information into binary-coded information. The ambel outputs are gated with the channel pulsed information. Only one channel pulse can propagate to the coding matrix. The coding is effected by diode coupling of the channel pulse to the proper set of binary buses. Figure 80 is a schematic diagram of the frequency-coding matrix for sub-bands 2 and 5.

e. CENTRAL CLOCK

176 Figure 81 shows the central clock, which consists essentially of four delay circuits. Two short-time circuits use delay lines and two long-time circuits use one-shot multivibrators. The first short-time-delay circuit clocks the frequency ambel. This clocking prevents the generation of erroneous frequency information due to the inequalities in the response times of the crystal-video receivers in the frequency analyzer. The second short-time-delay circuit clocks the output blocking oscillators of the frequency and direction analyzers. This second delay circuit ensures that the frequency and direction ambels operate properly before any information is taken from them to be applied to the computer.

177 The output from the first short-time-delay circuit is applied to a special one-shot multivibrator, which resets

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itself in a maximum time of 110 μ sec. The output from this multivibrator is applied via an OR gate to the r-f amplifiers in all sub-bands to prevent measurement of multiple signals while the width of one pulse is being measured by the pulse-width analyzer. Gating of the r-f amplifiers does not affect the pulse-width analyzer, because its information is extracted from the r-f system at a point ahead of the r-f amplifiers. At the end of the pulse-width analysis, the 110- μ sec multivibrator is reset by a signal from the pulse-width analyzer. If there is no reset signal received from the pulse-width analyzer (for example, when the central clock is triggered by a noise pulse), then the multivibrator automatically resets itself 110 μ sec after it is triggered.

178 The output from the 110- μ sec multivibrator is applied to a 25- μ sec multivibrator, triggering the 25- μ sec multivibrator on the trailing edge of the 110- μ sec pulse. Both multivibrator outputs are applied to the OR gate that controls blanking of the r-f amplifiers. The total time of r-f blanking is therefore 25 μ sec longer than the width of the received pulse. The additional 25- μ sec interval allows recovery of the video circuits of the various analyzers and prevents acceptance of any other pulses during this period.

7. PULSE-WIDTH ANALYZER

a. EARLY EFFORTS

179 The early efforts in the design of the pulse-width analyzer were directed toward the development of the 1-Mc pulse source, trigger circuits, and the 1-Mc counter stages. The 1-Mc pulse source used a crystal-controlled Clapp oscillator circuit and regenerative pulse shaper. The pulse source operated in a satisfactory manner; however, we decided to replace the basic free-running driver with a start-stop oscillator. A start-stop oscillator output is coherent with the leading edge of the pulse to be measured,

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thus providing greater accuracy in measurement than that afforded by a free-running driver (at the same clock rate).

180 The cascaded counter stages used in the early efforts were conventional flip-flops using diode steering. Serious difficulties were encountered at low temperatures, where the counters failed to operate. Modifications of the basic counter stage did not improve its operation, and we decided to suspend its use in favor of a circuit developed at Lincoln Laboratory. The final designs are described in paragraphs 181 through 193.

b. FINAL DESIGN

181 The pulse-width analyzer accepts as inputs both phases of a first-difference signal from the driver for the pulse-width analyzer (see Figure 66) and delivers as an output a 7-digit binary code indicating the pulse width. It also delivers an end-of-pulse output to the central clock to reset the 110- μ sec multivibrator. The final design of the breadboard of the pulse-width analyzer operates reliably over the -62 to +87 C temperature range, with power-supply variations of ± 10 percent. The accuracy is $\pm(0.5 \text{ percent} + 1 \mu\text{sec})$.

182 The data output of the pulse-width analyzer is delivered to the computer about 12 μ sec after the pulse width is measured. The analyzer is ready to perform a new analysis about 30 μ sec after the end of the previously analyzed pulse. The data output consists of 4- μ sec-wide pulses on seven output lines so that the pulse arrangement corresponds to the simple binary code for the pulse width in microseconds. Received pulses up to 126 μ sec in width are measured to the nearest microsecond. Longer pulses are arbitrarily indicated as having a length of 127 μ sec.

183 Figure 82 is a block diagram of the pulse-width analyzer. A pair of signals from the output phase splitter of the driver of the pulse-width analyzer is applied to

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All and A12 of the pulse-width analyzer. The outputs from A11 and A12 are +PW and -PW signals, respectively. The +PW signal is delayed 1 μ sec to shorten the count in the analyzer by this amount, because the pulse-width analyzer driver adds an average of 1 μ sec to the width of pulses passing through it. At the beginning of pulse-width analysis, the delayed +PW signal passes through gate G10, which is normally open, and sets flip-flop FF1. Flip-flop FF1 starts a start-stop 500-kc oscillator, and permits the output from the oscillator to be gated via gate G1 to the counter stages. Blocking oscillators B1 and B2 produce 0.1- μ sec pulses on alternate half-cycles of the 500-kc oscillator output. Blocking oscillators B1 and B2 alternately set and reset high-speed flip-flop FF3. In addition, B2 triggers the counter chain consisting of FF4 through FF9.

184 At the conclusion of the input pulse, the -PW signal resets FF1 through G11. This causes the start-stop oscillator to stop oscillating after a few cycles and immediately closes G1 so that no further counts can be recorded in the counter. Simultaneously, the -PW signal triggers monostable multivibrator FF10, which closes G10 for 25 μ sec, so that FF1 will not be triggered by the overshoot from the negative pulse of the +PW signal. The -PW signal, through G11, also initiates the read-out process. Blocking oscillator B13 produces a 12- μ sec pulse that, at its conclusion, triggers B14. The 4- μ sec pulse on B14 gates the outputs from FF3 through FF9, via G3 through G9, respectively; the gated outputs are amplified by A3 through A9, respectively, to provide the 4- μ sec output data pulses. The trailing edge of the output from B14 triggers B15, whose output resets all other counter flip-flops. The read-out process takes about 20 μ sec.

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185 In the event of an overflow--that is, a pulse longer than 126 μsec --FF9 will be set and then reset before the -PW signal will stop the count. The reset of FF9 will trigger B16, causing an input to G2. This action will normally occur for pulses between 63 and 127 μsec in width, when all flip-flops are reset by B15 after the read-out. In the event of triggering of B16 by a normal reset, FF1 will have been reset by the -PW signal 16 μsec before any possible output from B16; thus, G2 will be closed to the B16 pulse. In the event of an overflow, FF1 will not have been reset; thus, G2 will be open. The G2 output, amplified by A10, will reset FF1 and initiate the read-out process via G11. Simultaneously, the G2 output will (1) set all counter stages so that all ones will be read out, indicating a pulse duration of 127 μsec , and (2) set FF2, which, in turn, will close G10 to incoming +PW signals until FF2 is reset by B15 at the conclusion of the read-out process.

186 The output from G2 resets FF1, which, in turn, closes G2. Thus, if FF1 were connected directly to G2, the G2 output would be very narrow, equal to the sum of the delay through G2, A10, and G11, and the turnoff time of FF1, and the G2 output would be of insufficient width to trigger FF2 and B13. The FF1 output is delayed by a time longer than the width of B16 but less than 12 μsec ; thus, the output from G2 will be the full width of B16, which is more than ample to trigger FF2 and B13.

187 In the event that no -PW signal is received, the analyzer will be ready to receive +PW signals when FF2 is reset at the end of the read-out process. If a -PW signal occurs, FF10 will be triggered to prevent the +PW signal overshoot from triggering FF1; if B13 has recovered after being triggered by the overflow, it will initiate a read-out process. Since all flip-flops will have been reset, gates G3 through G9 will remain closed to B14, and there will be no output from the analyzer.

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188 High-speed flip-flops FF1 through FF4 are designed on the basis of symmetrical back-clamped non-saturating flip-flops that were developed at Lincoln Laboratory. Figure 83 is a schematic diagram of the pulse-width analyzer. The flip-flop consists of two RC-coupled flip-flops back to back, with collector loads in common. The collector load uses series peaking to provide additional current at the start of a change of state to satisfy the requirements of the speed-up capacitors, which are used in the NPN half of the flip-flop.

189 The pulse response of the 2N495's, however, is sufficiently fast that the PNP half of the flip-flop is run without speed-up capacitors. The 0.02- μf capacitors connecting the PNP and NPN bases are used in triggering the flip-flop. Since the base of the 2N118 is always negative and that of the 2N495 is always positive, a charge is stored in these capacitors. The capacitors are made sufficiently large that this charge remains substantially constant; thus, there is a substantially constant potential difference between the bases of the 2N118's and the 2N495's. This difference is such that when one transistor is conducting, the other is not.

190 To set FF3, the lower 2N338 is triggered by a narrow pulse. This causes its collector, and thus the bases of the right-hand 2N495 and 2N118, to drop in voltage, turning the 2N118 off and the 2N495 on. The action of the flip-flop collectors through the RC-coupling networks reinforces this action. A similar process occurs during reset.

191 Each of the flip-flop transistors has attached to its base a diode network whose function is to prevent saturation. If the diodes are of the same type and at the same temperature, the base and collector of a given transistor will tend to stay at the same potential (the potential at

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the junction of the two diodes plus the diode forward drop) when the transistor is turned on. If the collector voltage rises too high, it will cut off the diode to which it is connected, thus providing more base current, which will cause the collector voltage to drop. If the collector voltage drops, more current will be drawn through its diode, lowering the base current, which will cause the collector voltage to rise. Thus, the collector is maintained at the base voltage, which is sufficient to prevent saturation.

192 Medium-speed flip-flops FF2, FF5, FF6, and FF7 are conventional RC-coupled circuits with anti-saturation networks similar to those in the high-speed flip-flops. A resistor is used, instead of one of the diodes, in the anti-saturation networks, because some degree of saturation is permissible. Flip-flop FF5 has somewhat different component values from the others to ensure faster operation. The counter stages use conventional diode steering. Low-speed counter stages FF8 and FF9 are entirely conventional and also use diode steering.

193 The start-stop oscillator is essentially two emitter followers boot-strapped to a tank circuit. Oscillation begins when the current through the inductance is interrupted. Oscillation ceases when the current is turned on and the tank circuit is damped. With a tank circuit consisting of a video peaking coil, mica tank capacitors, and ceramic coupling capacitors, the frequency of oscillation in the breadboard circuit increased by 1 percent of the room-temperature value at both hot and cold temperature extremes. Emitter-follower coupling is used to eliminate loading effects on the output from the oscillator.

8. COMPUTER AND PRF ANALYZER

a. GENERAL REQUIREMENTS

194 The computer must (1) accept information from the frequency, direction, and pulse-width analyzers, (2) process

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the accepted information and determine PRF data, and (3) temporarily store all information until it is read out to the tape recorder.

195 Processing involves assembling all the data applicable to one intercept to an appropriate data group, or word. It also involves eliminating redundant data in the intervals between read-outs. Another processing operation is the rejection of intercepts that do not meet a minimum standard of repetition; the purpose of this rejection is to reduce the amount of spurious data that is produced by noise.

196 The computer is required to accept data at a random rate. The shortest time interval between a pair of intercepts from the same radar will be 100 μ sec and the longest will be 50,000 μ sec. A complete intercept word consists of the following basic data in binary digit form:

Frequency	7 bits
Direction	5 bits
Pulse width	7 bits
PRF	10 bits
Total	29 bits

197 The computer must be capable of storing these four pieces of radar information for each of as many as five different signals. For the present system, this requires a memory capacity of 29 bits in each of five storage units (called trunk memories), or 145 bits. One of the important considerations in the design of the computer is flexibility in the number of trunks and in the number of bits per signal that can be used. It is desired that this type of system be capable of use in equipments that are to have a different capacity from the equipment presently under consideration. This requires that the equipment be designed from the building-block point of view, where a complete equipment can be built by wiring together sufficient small standard blocks to equal

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the desired capacity. This applies not only to the memory but also to the other computer circuits.

198 To accomplish the data-processing functions of the computer, the frequency and direction memories must be interrogated each time that frequency and direction data are presented by the frequency and direction analyzers. This interrogation permits comparison of the new data with the data stored in the trunks; it must not destroy data already stored. Interrogation will occur at random times, the minimum interval between a pair of interrogations being about 25 μ sec.

199 Frequency and direction information will normally arrive at the computer simultaneously. This is accomplished through the action of the coincidence gate and allied circuits in the frequency and direction analyzers. Because these analyzers are threshold-operated devices, an occasional burst of noise will energize either analyzer or both. It is the function of the coincidence gate to eliminate spurious outputs from the analyzers arising from noise-generated signals in one of them. However, the coincidence gate cannot discriminate against spurious outputs arising from simultaneous noise-generated signals in both analyzers. To guard against recording these noise pulses, the computer is required to check that at least three successive frequency and direction pulses of the same binary code have been received within a time interval of 0.1 second. Where this condition is not satisfied, the intercept information that has been stored on this signal must be erased from the computer.

200 If the frequency and direction data at the input to the computer match those stored in any trunk, a trunk-match pulse is generated that is used to perform the following functions: (1) prevent the storing of the input information in any idle (empty) trunk; (2) perform in a timing circuit

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the anti-noise check described in paragraph 199; (3) provide the necessary coincidence for storing pulse-width data in the proper trunks; and (4) determine the PRF.

201 If the frequency and direction information at the input to the computer does not match that which is stored in any trunk, the data constitute a new intercept. It is then necessary to store this information in any empty trunk.

202 Pulse-width information is derived from a binary counter that begins counting at the leading edge of the received radar pulse and stops counting at the trailing edge. Pulse-width information thus becomes available at some time up to 127 μ sec after the frequency and direction information is received (since the widest pulse to be measured is 127 μ sec). Since the pulse-width counter is a threshold-operated device, it may be necessary to wait several radar pulse periods after frequency and direction information is received before pulse-width information is available. The computer, therefore, must be able to accept pulse-width information when it becomes available.

203 PRF information is determined by a binary counter that measures the time interval between pulses from the same radar. Because the pulses of several radars can be received during the pulse period of one radar, their pulses may interlace; this makes it necessary to identify pulses from the same radar to obtain a true PRF measurement. These pulses are identified by the trunk-match pulses, which are used to start and stop the PRF counter. (Although the trunk-match pulse will not be in time coincidence with the corresponding radar pulse, the spacing between match pulses will be the same as that between radar pulses.) Because the computer will store more than one signal, it is necessary to provide either one counter, together with the required selector circuits, for the entire computer, or a counter for each trunk.

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The individual stages of a binary counter are very similar to, and not much more complicated than, the flip-flop memory cells to be used for storage in the trunks. Therefore, an individual PRF counter was used in each trunk.

204 In summary, the computer must perform the following functions: (1) select an empty trunk in which to store the information corresponding to a particular intercept; (2) control the transfer of that information to a trunk memory; (3) prevent the same information from being stored in any other trunk; (4) verify that the stored information represents a legitimate intercept; and (5) retain the verified information until reset.

b. CHOICE OF COMPUTER TYPE

205 Two computer types, or data-handling methods, have been considered. One is the parallel method used in the AN/DLD-1(XA-1). The other is a series-parallel mode, wherein data are read in (in a parallel manner) and read out (in a series of parallel groups of bits) as required by the format chosen for the tape recorder. Internally, however, data are handled serially. (This computer is described in Appendix A.) The success of the parallel-type computer has been proven in the AN/DLD-1(XA-1). The series-parallel type of computer was investigated in the hope that it would provide a sizable reduction in the number of components; however, in a computer of the size required for the AN/DLD-2(XA-1), negligible component reduction is obtained. In a larger computer, the savings would become more significant.

206 In judging the relative merits of these two approaches, we have tried to view them in the light of the ferret problem as a whole. It is therefore necessary to consider the possibility of appreciably higher intercept rates than the AN/DLD-2(XA-1) is expected to handle. In addition, there is the ever-present problem of reliability. From both of these viewpoints, the parallel computer has the advantage in that:

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1. It is faster than the serial computer, as now planned; in the present state of the art, the parallel computer can be speeded up still further, but the serial computer probably cannot.
2. It is inherently less liable to total failure than the serial computer.

207 The three possible relationships between each intercept and the status of the computer when the frequency and direction data of the intercept are presented to the computer are as follows:

1. The intercept does not match any stored intercept, and at least one trunk is empty.
2. The intercept does not match any stored intercept, and all trunks are busy.
3. The intercept matches a stored intercept.

208 In the first case, the parallel computer will store the new intercept in less than 10 μ sec after it is presented; the serial computer will take nearly 15 μ sec. The speed given for the serial computer in this case is based upon the use of a 12-bit shift register, for frequency and direction together, operating at 10^6 pps. This rate would require more costly transistors than those that are needed in the parallel computer, and the shift-register stages would probably be more elaborate than the relatively slow-response flip-flop memories that the parallel computer uses.

209 In the second case, the parallel computer can be made to indicate this condition within about 5 μ sec after an intercept is presented to it. Although this facility is not used in the AN/DLD-2(XA-1), it is quite possible that a higher-speed system, operating against narrow-pulse radars, might use this facility to accelerate the recovery of the receivers so that the probability of receiving matching intercepts will be increased. The serial computer makes no distinction in processing time between this second case and the first case.

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210 In the third case, timing is the same as in the second case. Both types of computers use a match pulse to initiate read-in into the matched trunk of the data generated in the pulse-width analyzer.

211 Judged on these three cases, the parallel mode of data-handling presently enjoys a small but clear margin over the serial mode in operating speed. Furthermore, the speed of the parallel computer is more conservatively rated, and acceleration by a factor of two can probably be attained, if required. Since the computer designed for the AN/DLD-2(XA-1) may well be applied to higher-speed systems, such as a transistorized AN/DLD-1, we have decided to use the parallel mode of data-handling.

212 An additional factor in our decision is reliability. The serial mode of computation is, in some respects, more frugal in its use of circuit elements. It has already been shown that this frugality is bought at the price of a disadvantage in speed. There is a further disadvantage in this frugality, which is obtained through the use of proportionately more common circuitry than in a parallel computer. A serial computer is more likely to fail in ways that cause the loss of all intercepts than is a parallel computer.

c. CHOICE OF MEMORY CELLS

213 About one-half of the circuitry in the computer consists of temporary stores or memory cells. The choice of the memory cell thus becomes an important determinant of the size, weight, and power requirements of the computer.

214 A memory cell for the computer must meet these basic requirements:

1. The cell must be reliable--it must have long life under specified environmental conditions.
2. Future production and availability must be certain.
3. The read-in time must be less than 15 μ sec.

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215 These additional considerations must be weighed in choosing the most desirable memory cell:

1. The power drain should be low.
2. The cell should be small and light in weight.
3. The cell should be set and reset easily, and be easy to interrogate.
4. The cell should be capable of fast reset.
5. The cell state should be easy to read.
6. Visual indication of cell state should be easily provided.
7. The cost should be low.
8. The cell components should be available from more than one manufacturer.

216 Transistor flip-flops, unijunction transistors, pulse-driven magnetic cores, and r-f driven magnetic covers were studied for use as memory cells.

217 Families of parameters of unijunction transistors exhibited very wide spreads at room temperature. Temperature variations can be expected to superpose further large changes in the parameter spread. Further, there was only one manufacturer of unijunction transistors. For these reasons, application of unijunction transistors was not considered.

218 R-F driven cores (the Potter Magnistor) were not used, because (1) they are specialized devices manufactured by only one supplier; (2) they require high power consumption from a special r-f supply, with possible attendant shielding problems; (3) they are relatively large and heavy; and (4) they cannot easily provide indication of state.

219 A computer using pulse-driven magnetic cores and transistor drivers was compared with one using transistors only. From considerations of memory-cell requirements, reliability, general suitability to the intercept problem, development time, production, operation, and maintenance,

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a clear preference for the transistor-only configuration evolved. The basic flip-flop memory cell has eight components, excluding transistors, and draws about 60 milliwatts. Further, the transistor flip-flop meets the requirements described in paragraphs 214 and 215.

d. PARALLEL COMPUTER

(1) GENERAL

220 Figure 84 is a block diagram of the parallel computer. It differs principally from the computer of the AN/DLD-1(XA-1) (reference 1) in its empty-trunk finder and in the fact that PRF analysis is performed in each trunk. Furthermore, PRF is counted quasi-logarithmically. The new empty-trunk finder operates upon the trunks in parallel, whereas the finder of the AN/DLD-1(XA-1) searches serially; this change improves the speed of response of the computer, and obviates the necessity of regenerating a pulse as it is passed from trunk to trunk.

221 Figure 85 shows waveforms that are associated with the acquisition and storage of new frequency and direction data. Time-coincident frequency-data and direction-data pulses are presented to the computer by the frequency and direction analyzers; the coincidence of the pulses is ensured by the coincidence gate and associated circuitry in the direction analyzer. The direction pulses are sampled by an OR gate whose output starts the trunk supervisor by triggering the matcher-drive generator after a 3- μ sec delay. The frequency and direction pulses are also presented to the trunks; within each trunk, they are presented to the read-in AND gates of the appropriate memories and to the memory matcher. The memory-matcher circuits compare the incoming data with previously stored data, if any, to determine whether the incoming data represent a new intercept or a repeat of an intercept that has already been acquired. In the former case, the

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data are recorded in an empty trunk, if any, and in the latter case, the verification circuits in the trunk note that the data have been repeated.

222 The matcher-drive pulse, delayed 3 μ sec from the receipt of frequency and direction data, samples the states of the memory matchers. If there is a match in any trunk, the associated matcher emits a match pulse. The match pulse inhibits a gate that otherwise passes a trigger, delayed an additional 2 μ sec, to the enable generator. If there is no match pulse, an enable pulse is generated that samples the state of the empty-trunk finder. The empty-trunk finder is conditioned by busy, or occupancy, signals from each of the trunks, so that the enable pulse generates a trunk-gate (read-in) pulse that reads in the new frequency and direction data to the lowest-numbered unoccupied trunk. The trunk-gate pulse is slightly delayed relative to the enable pulse by the reaction time of the empty-trunk finder.

223 After frequency and direction are read into a trunk, the trunk's busy bus changes its state from "not occupied" to "occupied." When this happens, the voltage level of the bus falls. The busy inputs to the empty-trunk finder must not change the state until after the end of the enable pulse; overlap would cause the empty-trunk finder to respond as though the trunk that was just occupied had been busy from the beginning, thus causing a trunk-gate pulse to be emitted to the next higher-numbered unoccupied trunk.

224 The indicated time delays are conservative. If higher-speed operation than that required by the AN/DLD-2(XA-1) were desired, it is probable that, at the present state of the transistor art, the delays could be reduced by a factor of more than two.

225 As indicated, the trunk-match pulses will be used to instruct the pulse-width director on the proper routing

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of the data. Each receiver signal will be measured by the pulse-width analyzer. If the signal matches a signal that is stored in a trunk, its pulse-width data pulses will be routed to that trunk. If the pulse-width data have not been read into the appropriate memories, the memories will acquire these data at this time; if the data have been acquired, their repetition will not be allowed to affect the trunk.

226 The traffic-handling capacity of a computer can be considered in the light of two factors--instantaneous peak traffic density and average traffic density. The former factor influences the choice of computer type, because this factor is related to the response speed of the trunks and the supervisory circuits. Average traffic density influences the configuration of the computer--in particular, the number of trunks and the use of auxiliary circuits common to all trunks. The choice of the number of trunks is also influenced by the time over which traffic density is averaged--that is, the read-out period of the computer.

227 On the basis of available estimates of radar distributions and densities, and considering the ferret problem generally, we have chosen to use a five-trunk computer and a one-half-second read-out period. We have also chosen to analyze pulse width in a common circuit and PRF in each trunk individually. It is natural to analyze pulse width at the time that the pulse to be analyzed is being received; since the system has an instantaneous capacity of only one intercept, a common pulse-width analyzer does not act as a bottleneck. On the other hand, PRF is determined by counting the time between intercepts; in this interval, other signals can be received. To prevent a common PRF analyzer from becoming a bottleneck, the read-out interval of the computer should be several times the product of the number of trunks and the probable average reciprocal PRF.

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228 In addition to these basic criteria concerning pulse width and PRF, there are several others of a purely practical nature. One criterion is the rate at which each of the two analyzers must operate. The pulse-width analyzer operates at 1 Mc, the PRF analyzer at 250 kc. The driver for the PRF analyzer is a common circuit, even though the counters are repeated in each trunk: it is easier to distribute 250 kc than 1 Mc. Other criteria concern the types of circuits and transistors to be used. Counter circuits and memory flip-flops are very similar; for the present purpose, the former can serve as both. However, high-speed counters use appreciably more components than memories, draw more power, and require much more expensive transistors. These factors also favor the use of a common pulse-width analyzer (with associated memories in each trunk) and individual PRF counters.

(2) EMPTY-TRUNK FINDER

229 Figure 86 illustrates the principles of the empty-trunk finder. The signal indicating busy is positive (or the same polarity as that indicating enable), and the signal indicating not busy is negative (or the opposite polarity of that indicating enable). Since frequency and direction data must coincide, it is necessary to sample only the direction memories of a trunk to determine whether that trunk is busy. If the first trunk is not busy, this condition permits the enable pulse to produce a trunk-gate (read-in) pulse for the first trunk. Simultaneously, this condition prevents generation of trunk gates for the other trunks. If the first trunk is busy, the busy signal prevents generation of a trunk-gate pulse for that trunk; however, if the second trunk is not busy at the same time, then the combination of a busy signal from the first trunk and none from the second trunk permits the enable pulse to trigger the trunk-gate driver for the second trunk.

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These principles extend to the remaining trunks, and are summarized by the following Boolean equations, in which the subscripts 1 through 5 denote correspondence with the trunks that are so numbered, A_8 through E_{12} are direction memories corresponding to trunks 1 through 5, O is an occupancy (busy) signal, T is a trunk-gate pulse, and Z is the enable pulse.

$$O_1 = A_8 + A_9 + \dots A_{12} \quad (9)$$

and so on to

$$E_5 = E_8 + E_9 + \dots E_{12} \quad (10)$$

$$T_1 = ZO_1 \quad (11)$$

$$T_2 = ZO_1 O_2 \quad (12)$$

and so on to

$$T_5 = ZO_1 O_2 O_3 O_4 O_5 \quad (13)$$

An occupancy signal, O_n , is generated if any direction memory of the corresponding trunk is set. A trunk-gate pulse, T_n , is generated if simultaneously (1) the enable pulse Z is applied, (2) the n th trunk is not busy, and (3) all lower-numbered trunks are busy.

231 The change of state of the busy bus (after the initial read-in of frequency and direction) must not overlap the trunk-gate pulse. Symbolically, this would mean a change of O_n to O_n during T_n , in which case the equation for T_{n+1} would be satisfied. Since only T_n is desired, overlap is not permissible.

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(3) MATCHER

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The matcher is the circuit in each trunk that compares incoming frequency and direction data from the analyzers with the information already in the trunk. Incoming ONES are positive-going pulses, with their zero levels at ground, and incoming ZEROS are represented by the absence of pulses, with the input-line level remaining at ground. Stored ONES, on the contrary, are represented by a low d-c level (near ground), and stored ZEROS are represented by a high d-c level.

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Figure 87 is a schematic diagram of the matcher circuit. If two low levels are presented to a set of mixing diodes, the match bus will be held down, though the match-drive pulse tries to raise it and thus produce a trunk-match pulse. If two high levels are produced, they tend to allow the match-drive pulse to raise the match bus; however, these high levels raise the mismatch bus, turning on the match-inhibit transistor, and thereby hold the match bus down as in the other case of mismatch. If one high and one low level (the match condition) are produced by each of the pairs of data circuits to be matched, the mismatch bus will be down, holding the match-inhibit transistor off, and the match bus will be allowed to rise when the match-drive pulse is applied, thus producing a trunk-match pulse.

(4) PULSE-WIDTH DIRECTOR

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When an intercept is received that matches a previous intercept in frequency and direction, the trunk that already contains the intercept generates a trunk-match pulse. The match pulse, in addition to its other uses, is applied to a gate in the pulse-width director of its trunk. This gate has another input, which is capable of inhibiting propagation of the match pulse through the gate. The inhibit signal is generated by an OR gate that samples all of the pulse-width memories of the trunk. If any of the memories

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contains data, the match pulse is inhibited, and no further action is taken by the director. Therefore, in the absence of a match or the existence of previously stored pulse-width data, no pulse-width data are read in.

235 If a match pulse passes through the gate, it sets a flip-flop, which primes read-in AND gates via an emitter follower (which serves as a power amplifier). When pulse-width data are read out of the analyzer, they pass through the read-in gates and are stored in the memories. The memory OR-gate output then rises and, after a delay sufficient to permit the memories to settle, resets the flip-flop, thus closing the read-in AND gates. If there should be a pulse-width over-run, the over-run signal causes the flip-flop to reset. This happens before a read-out from the analyzer can take place, and thus prevents storage of erroneous data. Figure 88 is a block diagram of the pulse-width director and memories.

(5) PRF ANALYZER

236 Frequency and direction pulses initiate the data-handling process in the computer. When the data are newly read into a trunk, the PRF counter in that trunk is started. On the second match pulse, the PRF counter is stopped.

237 The PRF counter actually measures the pulse recurrence interval, rather than the frequency, by measuring the number of 4-μsec intervals between the first intercept (which starts a counter) of a signal and the third (which stops the counter). The choice of 4-μsec intervals is based upon the maximum PRF to be measured (10 kc), the required accuracy (±5 percent), and the number of intercept intervals during which the counter runs (two). The PRF counter is also used to verify intercepts; if the third (counter-stopping) intercept is not received within 100,000 μsec (two pulse periods at the minimum acceptable rate of 20 pps), the intercept

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data associated with the counter are considered invalid, and the trunk is cleared. If the third intercept is received within the maximum time, the data are considered valid, and a memory flip-flop is set to indicate this validity.

238 PRF counting is performed in a quasi-logarithmic fashion. This is accomplished by starting the counting process at a high rate (4-μsec intervals between drive pulses), and reducing the rate in steps at specified count levels as the total accumulated count increases. This approximates a true logarithmic counter, in which the count rate would decrease continuously as the total count increased.

239 Figure 89 is a block diagram of the quasi-logarithmic PRF analyzer. The basic counting rate is determined by the shortest interval (corresponding to the highest PRF) to be measured and the required accuracy. The highest PRF is 10 kc; the shortest interval then is $2 \times 1/10,000 = 200 \mu\text{sec}$, the factor 2 entering because the PRF counter is used for intercept verification. The required accuracy is ±2 percent. Thus, $0.02 \times 200 = 4 \mu\text{sec}$ is the basic counting interval, and $1/4 \times 10^{-6} = 0.25 \text{ Mc}$ is the counting rate.

240 The largest number to be counted is determined by the longest interval (corresponding to the lowest PRF) to be measured and the basic counting rate. The longest interval is $2 \times 1/20 = 100,000 \mu\text{sec}$ and the largest count is $0.25 \times 100,000 = 25,000$.

241 The quasi-logarithmic analyzer is divided into two groups of circuits--common and trunk. Common circuits serve all five trunks; these circuits consist of the basic rate generator and count-down circuits. Trunk circuits are those that are included individually in each trunk.

242 Since one set of drivers serves five sets of counters whose starting and stopping times occur at random,

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there is no coherence of the counters and their drivers. Thus, the operation of the quasi-logarithmic analyzer proceeds as follows.

243 The counter is started by the trunk gate, which sets a flip-flop. In the set position, this flip-flop primes an AND gate to admit drive pulses at the basic rate. When a count of 256 in the PRF register is reached, the flip-flop is reset, unpriming its AND gate. The reset change of state produces a pulse that sets a second flip-flop; the second flip-flop then primes an associated AND gate that admits drive pulses at 1/16 of the basic rate. The count proceeds for an additional count of 288 at 1/16 rate, 80 at 1/64 rate, and finally 111 at 1/128 rate. This requires ten counter stages corresponding to the largest number 1023.

244 As can be seen from Figure 89, the saving of five counter stages (25,000 would require 15 counter stages) is not free; appreciable auxiliary circuitry is required to obtain reduced counting-rate drives, and to connect the drives as required. The number of components required for quasi-logarithmic PRF analysis is almost the same as that required for linear PRF analysis. However, the reduced number of bits produces a saving in read-out circuitry and magnetic tape required for storage. The decrease should shorten the time required for ground data handling and also will reduce the bandwidth required by the radio data link.

245 If two match pulses are generated before the count of 512 is reached, they are counted down (in time) by a factor of 2, and they set a flip-flop that serves as the verified intercept memory (VIM). When the VIM is set, it resets the driver-control flip-flop that is set; this action stops the count. The VIM is also coupled to the read-out circuits. After the trunk is read out, a reset pulse from the read-out circuitry triggers the reset generator, clearing the trunk.

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246 Figure 90 is a schematic diagram of the computer supervisor, and Figure 91 is a schematic diagram of the computer trunk.

9. READ-OUT UNIT

247 The function of the read-out unit is to sense the data stored in the trunks of the computer and to transfer these data to the recording heads of a multichannel tape recorder. The read-out unit must connect the recorder heads to the trunks in accordance with a program suitable to the ground data-handling equipment; some logical rules of the system, such as the requirement that only verified intercepts are to be read out, influence the program. In addition to reading out the intercept data gathered by the AN/DLD-2(XA-1), the read-out unit must transfer data from a navigational-data digitizer to the tape recorder. The intercepts are thus correlated with navigational information on the tape.

248 The tape format is an important consideration in the design of the read-out unit. The tape format must be applicable to the requirements of the AN/DLD-1 as well as to the AN/DLD-2, and must use magnetic tape economically to minimize the bulk and weight of tape and reels required for a mission. Discussions were held with RADC relative to the ground data-handling problems and the development of a suitable tape format.

249 Figure 92 shows the recommended tape format for the AN/DLD-1 and AN/DLD-2. The five-trunk AN/DLD-2 computer is read out twice per second. Each intercept occupies two rows on the tape. The navigational data occupy ten rows on the tape; thus, the read-out unit handles the navigational data digitizer as though it were a five-trunk computer. The navigational data are read out once per second.

250 Figure 93 is a timing and block diagram of the read-out unit. It has been simplified somewhat by the

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elimination of some details (for example, in the timing chains). There are two timing operations performed in the read-out unit. One timing operation conforms to the requirement shown in Figure 92, that one block of data is read out each second. This timing is governed by a signal received at 1-second intervals from the navigation computer. The other mode of timing governs the programming of the read-out unit within each individual block of data. This timing is controlled by a free-running oscillator, which operates at a rate of about 25 cps. In addition to these two repeated modes of timing, an initial reset signal received only once, at the beginning of operation, places the read-out unit into its proper initial position. This signal, which is associated with the turn-on of the power supply, ensures that the read-out unit starts in a valid manner. The initial reset signal will be used in the same manner by the other data-handling circuits, such as the computer.

251 The recording process begins when a 1-second signal is received. The 1-second signal sets FF2 and FF6. When FF2 is set, it generates a level called "allow selector," which enables an AND gate so that timing pulse A1 can drive the trunk selector. The trunk selector, consisting of a binary counter of three stages and a crystal matrix, decodes the counter output to produce five sequential trunk-examining signals. The fifth signal enables an AND gate that permits timing pulse B4 to reset the trunk selector after the fifth trunk in a group of five has been examined. The same fifth signal enables another AND gate that permits timing pulse A4 to step a unit called the 5's counter, which is a group of circuits similar to the trunk selector. The 5's counter is arranged to count three groups of 5's. After three groups of 5's have been counted, the 5's counter is reset by a timing pulse (B4) operating through an AND gate that is enabled when the third group of 5's is counted. The same timing pulse (B4) that resets the 5's counter also resets FF2.

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252 This action occurs at the end of a block of data, and the read-out unit then awaits the next 1-second signal. The trunk selector and the 5's counter have been so designed that the read-out unit is readily adaptable to a variety of ferret systems, which may have a different number of trunks and a different read-out rate from the AN/DLD-2(XA-1).

253 When FF6 is set by the 1-second signal, it generates the allow navigation-trunk signal. The allow navigation-trunk signal enables five AND gates, which, in conjunction with signals from the trunk selector, sequentially produce five navigation-trunk selector signals. These five signals enable enable AND gates in the navigation computer so that the data stored therein can be read out. Since each trunk occupies two rows on the tape, it requires two timing signals for the complete read-out of one trunk. The two read-out timing signals are first A7 and then B3. Following pulse B3, pulse B4 is applied to AND gates that are associated with the navigation computer. These AND gates, in connection with the sequential navigation-trunk selector signals, produce sequential navigation-trunk clear signals. The clear signals clear the trunks that have been read out to prepare them to receive new navigation data.

254 After all five navigation trunks have been read out, the fifth trunk-selector signal, in conjunction with timing pulse B4, resets FF6. The output from FF6 then changes from allow navigation trunk to allow data trunk. The allow data-trunk signal permits read-out of the ferret computer trunks in a manner similar to that in the navigation converter. Reading out the computer, however, requires a refinement not required in reading out the navigation converter. This refinement is that only trunks containing verified data are to be read out. (Since the navigation converter operates in a cyclical manner, coherent with the read-out unit, the

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precaution of verification of navigation data is unnecessary.) There are five AND gates associated with the read-out function of the computer, in a manner analogous to those that produce the navigation-trunk selector signals. However, the five AND gates associated with the computer produce the verified-data-trunk selector signals only indirectly. This is done to ensure that the data contained in a trunk are verified before read-out begins and to prevent the possibility of a trunk being read out partially, should the data in the trunk be verified in the middle of the read-out cycle of that individual trunk.

255 This function is accomplished by generating the verified-data-trunk selector signals in flip-flops (FF7 through FF11). If the data in a specific computer trunk are verified, as indicated by the state of its VIM, the associated flip-flop that generates the selector signal for that trunk will be set by timing pulse A4 just before read-out pulses A7 and then B3 are generated. After each computer trunk is read out, it is reset by timing pulse B4 in the same manner that the navigation converter trunks are reset after being read out. In the case of the computer, timing pulse B4 also resets the flip-flop that generates the selector signal for that computer trunk. The sequential searching operation of read-out through the five computer trunks is performed twice in each data block, because the trunk selector counts to five three times within each block (the third group of five generated by the trunk selectors is that used for read-out of the navigation converter).

256 Figure 92 shows that track 1 of the tape is used as a synchronizing track. During the read-out interval, the tape contains a one in each row, except the first. In the first row, track 1 contains a zero, whereas the remaining 15 tracks contain ones. Timing pulses A6 and B2, which are

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in coincidence with read-out pulses A7 and B3, respectively, insert ones into track 1 as long as the allow selector signal is present. The insertions of ones in track 1 are accomplished by a combination of one OR and two AND gates associated with track 1. The navigational-trunk-1 selector signal, which is inverted for this purpose, blanks track 1 in the first row.

257 A one is also inserted into track 2 in the first row through the use of this inverted signal. This is accomplished in connection with read-in pulse A7; the AND gates associated with the data tracks produce ones when negative-going signals are applied to them. In the first row, ones are similarly inserted into the remaining 14 data tracks, except that the navigation-trunk-1 selector signal is re-inverted and applied to inverting OR gates before being applied to the data-track AND gates.

258 The navigation-trunk and verified-data-trunk selector signals are applied sequentially to AND gates, located in the navigation converter and in the computer, respectively. There are many such gates, one for each data bit stored in each trunk; for simplicity, the gates have been represented by a single AND gate for each trunk. The stored data are fed through the AND gates at the appropriate times, and combined in 29 OR gates, corresponding to the 29 different data bits that can be stored in the computer trunks. These data are then read onto the tape by means of the data-track AND gates; 29 bits are compressed into 15 tracks by handling them in two groups, one consisting of 14 bits and the other consisting of 15 bits.

259 Figure 94 is a schematic diagram of the control circuits of the read-out unit. Figure 95 is a schematic diagram of the 5's counter and the record gates of the read-out unit. Figure 96 is a schematic diagram of the trunk gates of the read-out unit. The composite of these circuits

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represents the complete read-out unit as used in the feasibility demonstration.

10. POWER SUPPLY

Several series regulator designs were investigated for application to the power-supply problem. The essential purpose of the investigation was to develop the basic regulator configuration that would be adapted to the design of the system power supply.

The simplest configuration used a power transistor as an emitter follower. The transistor base was clamped by an avalanche diode. The regulator circuit had relatively low efficiency (about 55 percent) and operated about as expected.

A 2N389 silicon power transistor was used as the variable series resistor in a second regulator design. Although the nominal β is 30, sufficient loop gain was provided by a pair of low-power silicon transistors in an amplifier cascade. Again, avalanche diodes provided the voltage reference. The regulator operated satisfactorily over a temperature range of -50 to +60 C temperature range, for all values of load. At rated load, the regulator had an efficiency of about 60 percent. The principal component of temperature sensitivity was that of the associated voltage reference diodes.

The design of future regulator circuits appears straightforward. Silicon power transistors are becoming available with far better characteristics than those available with the 2N389 transistor. No serious problems are anticipated in the design of the transistorized power supplies for future AN/DLD-2 equipments.

Commercial power supplies were used in the feasibility demonstration of the AN/DLD-2(XA-1).

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11. SPECIAL TESTS AND TEST EQUIPMENT

a. JUNCTION-TRANSISTOR TESTER

At the start of the development program, there was no commercially available junction-transistor tester that would meet our requirements. Thus, we decided to build a simple, portable, battery-operated transistor tester. The instrument is capable of precise d-c measurements of collector-to-base current gain (β), collector cutoff current (I_{CO}), emitter cutoff current (I_{EO}), and collector-to-emitter saturation voltage (V_{sat} --not a standard symbol). These are parameters of special interest in switching-circuit design. The first three measurements are made at voltages from 0 to 15 volts in steps of 1-1/2 volts, and the fourth measurement can be made at collector currents from 0 to 5 ma in steps of 1/2 ma.

A simple and accurate method of measuring by an incremental d-c method is the basis of the instrument. The base current in a grounded emitter circuit is adjusted to provide the desired collector-current operating point. A precise increment of base current is removed, and the resultant decrease in collector current is noted. The collector-current meter is calibrated to read β directly. The measurements of I_{CO} , I_{EO} , and V_{sat} , plus a battery test position, are added by means of appropriate switching.

Figure 97 is a schematic diagram of the junction-transistor tester. The meter is a standard 2-percent 50-ma meter. The use of 1-percent resistors ensures that β and V_{sat} measurements are generally accurate to better than 3 percent. β measurements using the tester were in close agreement with those obtained by small-signal a-c methods.

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b. TEMPERATURE EFFECTS ON TRANSISTORS AND PARAMETER MEASUREMENTS ON TRANSISTORS AND DIODES

268 Short-circuit current gain is the transistor parameter that is expected to most affect the operation of the AN/DLD-2(XA-1) over the -55 to +71 C ambient temperature range. Since this assumes that silicon transistors are used, the collector cutoff current (I_{CO}) can be neglected. It is also assumed that the transistor is not damaged by repeated temperature cycling.

269 Texas Instruments, Inc. gives the same curve of the variation of the quantity $1 - \alpha$ (α = collector-to-emitter short-circuit current gain) with ambient temperature for several of its silicon transistor types. It was believed, however, that the experimental checking of this curve could give information on

1. The possible spread of curves for various transistors of the same type and of different types.
2. Whether the same curve applied to other silicon transistor types.
3. How the low-temperature portion of the curve for silicon compared with that for germanium transistors.

270 In reference to the comparison between silicon and germanium transistors, the curves supplied by Texas Instruments, for some of their germanium switching transistors indicate little difference in current gains at low temperatures between germanium and silicon transistors. Nevertheless, since there is a prevailing opinion that silicon transistors are inferior to germanium transistors at low temperatures, it has been considered desirable to clarify this problem.

271 When measurements of current gain were being made, it was convenient to measure saturation voltage on the transistor tester used. The latter measurement is the collector-

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to-emitter voltage in a common-emitter circuit having a base current equal to 0.25 ma and a collector current equal to 1 ma. Measurement of saturation voltage at this one fixed operating point is probably indicative of the relative performance in switching circuits of saturated transistors at other operating points.

272 Measurements were made on a group of twelve transistors of six different types (two of each type) using a switching panel with an ultimate capacity of 36 transistors. The different transistor types are specified in Table III.

TABLE III
TRANSISTOR INFORMATION

Type	Manufacturer	Material	Polarity
925*	Texas Instruments, Inc.	Silicon	NPN
926*	Texas Instruments, Inc.	Silicon	NPN
951	Texas Instruments, Inc.	Silicon	NPN
ZJ-12	General Electric Co.	Silicon	NPN
2N167	General Electric Co.	Germanium	NPN
2N123	General Electric Co.	Germanium	PNP

* Tetrode types measured with second-base leads disconnected.

273 A chassis, which contained the twelve transistors plugged into sockets, was located in a test chamber, and the 36 connections to the transistors were brought out through the switching panel to the transistor tester. The test chamber was cycled from room temperature up to about +93 C, back to room temperature, down to about -57 C, and back to room temperature.

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Figure 98 shows the spread of the normalized collector-to-base current gain ($\beta/P_{26.5\text{ C}}$) vs temperature for the eight silicon transistors. A theoretical curve is also shown for silicon transistors calculated from the Texas Instruments data sheets. From Figure 98, it can be concluded that this theoretical and nearly linear curve holds approximately for all the silicon transistor types tested, with a spread of about ± 30 percent at low temperatures changing to ± 60 , -10 percent at high temperatures.

275

Although these results are for a small sample, the distribution of curves within the extreme spread indicates that the results should hold approximately for a larger sample. Note that the results for the four germanium transistors at low temperatures are within this same spread. Thus, it can be concluded that the low-temperature variations of short-circuit current gain for silicon and germanium transistors are almost equivalent.* This bears out the Texas Instruments data.

276

Figure 99 shows the resultant spread of the normalized saturation voltage ($V_{26.5\text{ C}}$) for the silicon transistors. Again, the variation is similar for the germanium transistors (though the absolute magnitude of saturation voltage is generally much less for germanium transistors). It is also interesting to note that there is a nearly linear increase of saturation voltage as the temperature increases, in spite of the corresponding increase in β .

* Current gain of the germanium transistors could not be measured at the highest temperatures because of operating-point difficulties. As far as could be determined, however, the equivalence of temperature variation noted at low temperatures remains valid for temperatures up to 60 C .

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Perhaps the most important low-frequency parameter of transistors in the design of common-emitter stages is β . We plan to use many type 903 and 904 transistors; thus, we had available a sufficient sample to make a small-scale temperature test. The variation of β with temperature was measured on a group of twelve 903's and a group of twelve 904's. The data, normalized to 10 C , are shown in Figures 100 and 101. The absolute values, as well as the normalized values of β , exhibited a much smaller spread for the 904's than for the 903's.

278

Among other transistor parameters important to the designer of a flip-flop are the collector and base saturation voltages. These voltages will appear between the emitter and the electrode whose parameter is being measured when specified currents flow between the emitter and each of the other electrodes. Since these voltages determine the steady-state characteristics of a flip-flop, their variation from transistor to transistor and with temperature are very important.

279

We have chosen to operate our flip-flops at nominal saturation currents of $I_C = 2.56\text{ ma}$ and $I_B = 0.4\text{ ma}$. We have therefore measured saturation voltage for these currents as well as those stated in the published specifications. Our measurements include base saturation voltages as well as collector saturation voltages, and showed good uniformity from transistor to transistor and with temperature variations. Figures 102 and 103 show base saturation voltage and collector saturation voltage vs temperature for twelve 2N118 transistors. Figure 104 is a schematic diagram of the test circuit used for these measurements. When the switch (shown in Figure 104) is in position 1, saturation voltages measured in accordance with the published specification. When the switch is in position 2, saturation voltage is measured in accordance with the operating conditions of the flip-flop circuit shown in Figure 105.

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280 We have measured the voltage drop, at forward currents ranging from 0.05 to 5.0 ma, across ten type 1N625 diodes. These measurements were made at room temperature and are uniform from diode to diode (see Figure 106).

c. TESTER FOR VIDEO-DETECTOR CRYSTALS

281 A tester for crystals was designed and built to conveniently standardize the testing of crystals as video detectors. This tester, a bridge-type null device, eliminates the human error involved in determining tangential sensitivity from oscilloscope measurements. With a forward d-c bias of 10 μ a or more (used to improve sensitivity), 90 percent of the units tested gave tangential sensitivities within ± 1.5 db of that measured on an oscilloscope. Since tangential measurement on an oscilloscope is subjective, this accuracy is regarded as excellent.

282 The tangential sensitivity of a video-detector crystal in decibels can be expressed as

$$S = C + 10 \log_{10} FM \quad (14)$$

where

S = the tangential sensitivity
C = a constant determined by experiment for the crystal type under measurement
FM = the figure of merit of the crystal

also,

$$FM = \frac{\beta R_V}{\sqrt{R_V + R_A}}$$

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where

β = the power sensitivity
 R_V = the video resistance of the crystal
 R_A = the effective noise resistance of the test amplifier

The crystal tester determines the power sensitivity from the d-c conductance G_{dc} and from R_V , which can be measured with a small a-c voltage or a small d-c increment.

283 The forward current in a diode can be expressed as

$$I = KE^x \quad (15)$$

The d-c conductance is then

$$G_{dc} = 1/R_{dc} = I/E = KE^{x-1}$$

The a-c (video) conductance is defined as

$$G_V = dI/dE$$

Taking the derivative,

$$G_V = KxE^{x-1} = xG_{dc} \quad (16)$$

and

$$x = R_{dc}/R_V$$

284 Now, let a d-c increment or a small a-c signal be added to the diode voltage so that the peak excursions are

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$$E_{\max} = E + \epsilon/2$$

$$E_{\min} = E - \epsilon/2$$

Then, the effective current is given by

$$\begin{aligned} \bar{I} &= K/2 \left[(E + \epsilon/2)^x - (E - \epsilon/2)^x \right] \\ &\approx KE^x + (Kx/8) (x-1)E^{x-2}\epsilon^2 \end{aligned}$$

285

Now

$$\begin{aligned} \Delta I &= \bar{I} - I \\ &= (Kx/8) (x-1)E^{x-2}\epsilon^2 \\ &= \frac{g_v(x-1)\epsilon^2}{8E} \end{aligned}$$

The increment of power causing this increment of current is

$$\Delta P = g_v \epsilon^2 / 4$$

286

Therefore, the power sensitivity is given by

$$\begin{aligned} \beta &= \Delta I / \Delta P \\ &= \frac{g_v(x-1)\epsilon^2}{8E} \cdot \frac{4}{g_v\epsilon^2} \\ &= \frac{x-1}{2E} \end{aligned}$$

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But $E = IR_{dc}$; so

$$\beta = \frac{x-1}{2IR_{dc}}$$

and the figure of merit becomes

$$FM = \frac{x-1}{2IR_{dc}} \cdot \frac{R_v}{\sqrt{R_v + R_A}} \quad (17)$$

287

Figure 107 is a schematic diagram of the tester for the video-detector crystals. In this circuit, a forward bias current is supplied and adjusted by R5 (which has a very high resistance); the combination of BT1 and R5 acts as a current generator. With S1 closed and S2 open, the desired bias is applied. When the bridge is balanced, the resistance of R3 equals the d-c forward resistance of the diode.

288

A small increment of current is applied through the diode by BT2 and R6 when S2 is closed. Bridge balance is restored by adjusting R8, which applies a fraction of α of the total voltage E_2 (from BT2) to R3 through R7. (Let R_3 , R_6 , R_7 , and R_8 denote the resistances of R3, R6, R7, and R8.) At balance, no incremental current flows through R1 and R2, and the following conditions apply:

$$I = E_2/R_6, \quad \epsilon = \Delta E = \alpha E_2 R_3/R_7, \quad \text{since } R_3 \ll R_7$$

$$R_v = \Delta E / \Delta I = \alpha R_3 R_6 / R_7 = \alpha R_3, \quad \text{since } R_6 = R_7$$

Now, $R_{dc} = R_3$, so $\alpha = R_v / R_{dc}$. From equation 16, we then see that $x = 1/\alpha$.

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289 For the case where the noise contributed by the amplifier is negligible, $R_A \ll R_V$, and the figure of merit is given by

$$\begin{aligned} FM &= \frac{(x-1)\sqrt{R_V}}{2IR_{dc}} = \frac{x-1}{R_{dc}\sqrt{R_V}} \cdot \frac{1}{2I\sqrt{R_V}} \\ &= \frac{1/\alpha - 1}{1/\alpha} \cdot \frac{1}{2I\sqrt{R_3}} = \frac{1-\alpha}{\alpha} \cdot \frac{1}{2I\sqrt{R_3}} \end{aligned} \quad (17')$$

In terms of α , R_3 , and bias current, tangential sensitivity in db becomes

$$S = C + 10 \log_{10} \frac{1-\alpha}{2I\sqrt{\alpha R_3}} \quad (14')$$

where C is experimentally determined for each crystal type. A nomograph in the form of a circular slide rule was constructed to quickly compute the tangential sensitivity from the easily measured quantities I , α , and R_3 .

d. PULSE-TRAIN GENERATOR

290 Figure 108 is a block diagram of a pulse-train generator. This instrument was developed primarily to simplify making recovery-time tests of video amplifiers. However, it is sufficiently versatile to have many other uses. A free-running (astable) multivibrator serves as the time base of the unit. When the CONTINUOUS-SCAN switch is in the open (CONTINUOUS) position, the AND gate after the astable multivibrator is always open, and the unit generates a continuous train of pulses. When the switch is in the closed (SCAN) position, the ring counter, cooperating with the two monostable multivibrators that follow it, opens the AND gate for a sufficient interval to permit three time-base pulses to

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pass. The AND gate is then closed for a time determined by the cascaded delay multivibrators, after which it opens again for three pulses, and then the cycle repeats. In this way, bursts of output pulses are generated that simulate pulses received from a scanning radar.

291 Each channel of the generator is capable of producing pulses whose width is variable from 1 to 100 μ sec; the controls of the separate channels are independent. The output from channel 2 can be delayed with respect to that of channel 1; and the output from channel 3 can be delayed with respect to that of either channel 2 or channel 1, depending upon the setting of the SERIES-PARALLEL CHANNEL DRIVE switch. When this switch is in the SERIES position, the output from channel 3 can be delayed up to 200 μ sec after the occurrence of the leading edge of the output pulse from channel 1. When the switch is in the PARALLEL position, the output from channel 3 can lead or lag the output from channel 2, with continuous control over this relationship. The outputs from the three channels are available separately or mixed in an OR gate; both output modes can be used simultaneously. The output amplitudes of the individual channels have independent controls.

292 The pulse-train generator is fully transistorized. Since the generator is intended for laboratory test use, germanium transistors are used.

e. TEST-PROBLEM GENERATOR

293 Figure 109 is a block diagram and Figure 110 is a timing diagram of the test-problem generator. The generator produces trains of pulses that can be programmed to simulate intercept data from a number of radars. Figure 109 shows that six radar intercepts can be simulated. The purpose of the extra intercept word (one more than the number of trunks in the computer) is to simulate a condition of high intercept activity when intercepts arrive faster than they are read out.

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294 The generator is driven from a variable-frequency oscillator and can run at a rate as high as 67 kc. This rate corresponds to a time between intercepts of 15 μ sec, which is less than the present recovery time of the video amplifiers of the frequency and direction analyzers. The highest rate at which the computer can operate can thus be checked. The sine-wave input is squared, and the square wave drives a 4- μ sec blocking-oscillator pulser and a five-stage binary counter. The pulser and the first stage of the counter are triggered on alternate half-cycles of the square wave; thus, their outputs are interleaved in time. The outputs from the counter are fed to a crystal matrix, which is also driven by the pulser.

295 The pulser clocks the signals selected by the matrix so that the output lines of the matrix carry steep-sided 4- μ sec pulses. The pulse intervals are determined by the counter and the interval gating connections of the matrix (see Figure 110). The matrix outputs are interconnected to provide simulated system signals, as shown in Figure 109, with some variations in the test program easily produced by means of switches.

296 At time 0, all trunks are reset (see Figure 110); the computer is thus placed in its initial position, with all trunks ready to accept intercepts. At time 1, simulated radar 1 is intercepted; its frequency and direction data are determined by the switches connected to the radar line. At time 2, simulated radar 2 is intercepted. At time 3, radar 1 is repeated, as it is again at times 5, 7, and 12. Radar 2 is repeated at time 13; the other simulated radars also are intercepted more than once each. At time 10, any one or more of the trunks can be reset in response to the setting of switches provided for this purpose.

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297 Repetition of the intercepts tests the matchers (the matcher circuit is the unit in each trunk that compares each intercept, as it arrives, with previously stored frequency and direction data). When the PRF and pulse-width analyzers are interconnected with the computer, this test of the matchers will also serve to test the analyzers. The multiplicity of radars tests the empty-trunk finder; an additional test of the empty-trunk finder is made when one or more of the trunks is reset at time 10. Since radar 6 is intercepted at times 9, 11, and 17, the empty-trunk finder should cause it to be lost at time 9 (since all trunks are full at this time), but at time 11, if a trunk has been reset at time 10, radar 6 should be read into the newly available trunk. A further check on the empty-trunk finder is made if more than one trunk is reset at time 10; in that case, proper operation requires that radar 6 be read into the lowest-numbered empty trunk, and that none of the radars whose data are already stored be read into an empty trunk.

12. FEASIBILITY DEMONSTRATION

a. BASIC PLANNING

298 Late in May 1957, at the request of the WADC project engineer, we submitted a proposal for expanding the research phase of this project. This expansion is intended to lead to a breadboard system capable of proving the feasibility of the techniques used in the AN/DLD-2(XA-1). Early in July 1957, we received the oral authorization of the WADC project engineer to proceed in accordance with our proposal.

299 In our proposal, we repeated our understanding with WADC that the AN/DLD-2(XA-1) would neither include means for determining the polarization of received signals nor provide channelized direction finding in sub-band 1; however, a left-right indication would be provided in sub-band 1. We also stated our belief that the airborne data-link encoder and ground-based decoder, to be used for telemetering of

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reconnaissance data from air to ground, are auxiliary to the basic ferret system. As such, they are not necessary to a demonstration of feasibility of the ferret approach, and we therefore proposed that they be eliminated from the feasibility demonstration.

300 The feasibility of AN/DLD-2 techniques was demonstrated by breadboarding an essentially complete system operating in two of the five sub-bands (2 and 5). Thus, the most difficult r-f problems (particularly those associated with d-f instrumentation at the highest frequencies) and the most difficult antenna problems at the lowest frequencies were combined. The antennas for sub-band 5 were demonstrated in a full-scale mock-up of a section of a B-47 aircraft, and the antennas for sub-band 2 were demonstrated in a 1/8-scale model of the complete B-47. The demonstration system included all video, data-handling, and read-out circuits. In addition to the basic demonstration system, of course, all components unique to each of the three other sub-bands were breadboarded, including antennas mounted on scale-model aircraft.

301 The basis for feasibility proof was a laboratory demonstration of the basic breadboard system, including use of the aircraft mock-up and model, together with a laboratory demonstration of the unique components of the three sub-bands not covered in the basic breadboard system.

b. DEMONSTRATION MODEL

302 Figure 111 is a block diagram of the sub-band 5 portion of the proposed feasibility-demonstration system. The antennas for sub-band 5 were mounted on a full-scale mock-up of a section of the B-47 aircraft. The components unique to this sub-band were fully breadboarded. The filter bank of the frequency receiver was complete. Actually, sub-band 2 and sub-band 5 will share video amplifiers, because these sub-bands are demonstrated simultaneously.

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303 Figure 112 is a block diagram of the sub-band 2 portion of the proposed feasibility-demonstration system. As in sub-band 5, the components that are unique to this sub-band are fully breadboarded. The antennas for sub-band 2 (1/8 scale) are mounted on the 1/8 scale model of the B-47 aircraft. The antennas for sub-band 2 will be tested with sub-band 5 instrumentation. Sub-band 2, exclusive of antennas, will be demonstrated with r-f signal generators. As shown in Figure 112, sub-band 2 will incorporate, for demonstration, means for suppressing interference from television signals.

304 Figure 113 is a block diagram of the portion of the feasibility-demonstration system that is common to sub-bands 2 and 5, as well as to the complete system. This portion of the system incorporates all circuits common to frequency and direction analyzers--ambels, coding matrices, log video amplifiers, minimum-amplitude selector, and output blocking oscillators--and a complete pulse-width analyzer. The computer is complete for two-trunk operation, including PRF analysis.

305 Skeleton breadboarding of three additional trunks is provided to demonstrate proper operation of the computer supervisory circuits. A full read-out unit is provided; however, a visual neon-lamp indicator, instead of a magnetic tape recorder, is used as the output device. The recorder does not require techniques development, and visual indication is a more convenient means for monitoring system output. Provision was made in the read-out unit to slow down the rate of acquiring data, so that the detailed system operation could be conveniently viewed. Commercial power supplies were used throughout.

306 Figure 114 is a general view of the feasibility-demonstration system. Figure 115 shows the neon-lamp display for read-out of the feasibility-demonstration system.

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Figure 116 is a system tuning diagram, showing the progress of received signals from the input to the neon-lamp-display output.

307 A test-problem generator was built to simulate computer input signals. This test-problem generator, which is quite similar to that described in paragraphs 293 through 297, provides a convenient means of demonstrating the computer and read-out functions. Figure 117 is a block diagram of the test-problem generator for the feasibility-demonstration system, Figure 118 is a partial schematic diagram of the neon-lamp display, Figure 119 is a schematic diagram of the test-problem generator, and Figure 120 is a schematic diagram of the test-problem controls.

c. FEASIBILITY DEMONSTRATION OF COMPONENTS NOT INCLUDED IN SYSTEM MODEL

308 Since only two sub-bands were included in the bread-board model of the feasibility-demonstration system, it was necessary to prove the feasibility of techniques in the three other sub-bands by component breadboarding. This was accomplished as follows.

(1) ANTENNAS

309 Aircraft model measurements of patterns were made for a single pair of antennas in sub-bands 1, 3, and 4. A comparison of these patterns with complete results obtained in the two other sub-bands is sufficient to predict performances.

(2) R-F FILTERS FOR FREQUENCY ANALYSIS

310 Each of the 27 filters not incorporated in the over-all system were tested individually; in some instances, a single filter was used to cover more than one channel (by retuning). Paralleling of more than one filter across the feed line of an antenna was tested, where necessary.

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(3) R-F COMPONENTS FOR DIRECTION ANALYSIS

311 Each unique r-f component used for direction analysis in sub-bands 3 and 4 was tested. These components include hybrid junctions, r-f amplifiers, tapped lines, and broad-band crystal detectors. Components for left-right indication for sub-band 1 were also breadboarded and tested individually.

d. DEMONSTRATION PROGRAM

312 The following is a description of the procedures followed in the feasibility demonstration of the AN/DLD-2 system. The procedures demonstrate the functions and capabilities of the system.

313 The tests were performed on a partial system that incorporated circuits of every type used in the final system but with less duplication than required to complete the over-all system. Frequency measurement was demonstrated in two sub-bands (60 to 120 Mc and 480 to 1000 Mc) of the five planned to cover the 30 to 1000 Mc range. Direction-of-arrival measurements were demonstrated in the same two sub-bands (60 to 120 Mc and 480 to 1000 Mc), which represent the most difficult antenna siting problem and direction-finding microwave instrumentation problem, respectively.

314 Complete processing circuitry for frequency and direction data was demonstrated. Complete pulse-measuring equipment was also demonstrated. Temporary storage in the computer was in two full memory trunks and three skeleton trunks instead of five full memory trunks. Circuitry for PRF measurement was demonstrated in both full trunks.

315 Complete read-out circuitry was demonstrated. To make output information readily available, a neon-lamp display instead of a tape recorder was used as an output indicator. During the tests, at least one of the following input conditions was used:

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- Condition 1. Radio transmission from a test antenna to the sub-band 5 antennas, mounted in a mocked-up section of the aircraft, was made. The frequency was preset, as desired, before the tests; however, because of the relatively long time required to change the frequency of the AIL Type 124 power oscillator, it was not changed during a test run. The AIL Type 124 power oscillator was pulse-modulated from an external source.
- Condition 2. Radio transmission from a test antenna to the sub-band 2 scaled antennas, mounted in a 1/8-scale model of the B-47 aircraft, was made. Operation was at frequencies of 480 to 960 Mc. Here again, the frequency was preset and unchanged during a test run. The AIL Type 124 power oscillator was pulse-modulated from an external source.
- Condition 3. An r-f signal generator was connected to the antenna terminals of sub-band 5 through a hybrid that divides the signal equally and inverts one output to give a 180-degree phase shift. A fixed or variable length of line was added to one signal path for those measurements that require direction simulation. The signal generator was externally modulated.
- Condition 4. An r-f signal generator was connected to the antenna terminals of sub-band 2 through a hybrid that divides the signal equally and inverts one output to give a 180-degree phase shift. A fixed or variable length of line was added to one signal path for those measurements that require direction simulation.
- Condition 5. The test-problem generator was connected to the blocking oscillators at the output from the frequency-channel video amplifiers and the output from the minimum-amplitude selector (direction) channels. From one to six different radars can be simultaneously set up on the test-problem generator, giving any direction and any frequency channel.

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(1) INFORMATION MEASURED BY THE SYSTEM(a) DIRECTION

316 With only the direction-channel lamps energized and with input condition 1, it was demonstrated that the direction was measured. With repetitive pulse triggering of the signal generator, the neon indicators for direction-channel outputs showed instantaneous direction as the full-size mock-up was rotated.

(b) FREQUENCY

317 With only the frequency-channel lamps energized and with input condition 4, it was demonstrated that the frequency was measured. With repetitive pulse-triggering of the signal generator, the neon indicators for frequency-channel outputs showed instantaneous frequency as the signal generator was tuned.

(c) PULSE WIDTH

318 With only the pulse-width lights energized and with input condition 4, it was demonstrated that the pulse width was measured. With repetitive pulse-triggering of the signal generator, the pulse width was changed. The pulse-width indicators showed an instantaneous measurement of the pulse width in microseconds. The measurement was indicated in binary code; however, pulse width in microseconds can be determined by adding the numbers above each lighted neon lamp.

(d) PULSE REPETITION FREQUENCY

319 With only the PRF lamps energized and with input condition 4, it was demonstrated that the PRF was measured. With repetitive pulse-triggering of the signal generator, the PRF indicators showed the PRF. The reading shown on the lamps was a binary indication of a quasi-logarithmic count

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of the time interval between pulses. The actual PRF can be read from a conversion table (see Appendix B).

(2) BASIC SIGNAL-PROCESSING PRINCIPLES

(a) FREQUENCY AND DIRECTION

320 To explain reception and storage of frequency and direction, the pulse width, PRF, and read-out were turned off. Using input condition 5, one radar was set up and the pulser was pushed once. In the frequency analyzer, the frequency-channel lamp, corresponding to the frequency on the test-problem generator, flashed once. In the direction analyzer, the direction-channel lamp corresponding to the direction angle between the aircraft and radiating antenna, flashed once. The frequency- and direction-channel numbers were encoded into binary numbers and the corresponding lamps in the output blocking oscillators flashed once.

321 The coincidence lamp in the computer common flashed once. The computer binary-driver lamps flashed once, corresponding to the output blocking-oscillator lamps. The match-drive lamp flashed once. The acquire lamp flashed once, and the trunk read-in lamp in trunk 1 flashed once. The memory lamps in trunk 1, corresponding to the computer binary drivers, lit and remained lit. The busy lamp of trunk 1 lit. The pulser was pulsed again. The same lamps that flashed previously flashed again. Since information is stored in the trunk, the trunk match lamp flashed once. This inhibited the acquire and there was no flash on the acquire or on any trunk read-in lamp.

322 The video pulser was pulsed again. This simulated the receipt of the third intercept from the same radar. The same lamps that flashed for the second intercept, flashed for the third intercept. In addition, the VIM lamp lit and remained lit, signifying a verified intercept.

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(b) PULSE WIDTH

323 The trunk was reset with the trunk reset pushbutton. The pulse-width analyzer was turned on and the video pulser was pulsed once. Frequency and direction were stored in trunk 1. Pulse width was measured in the pulse-width analyzer. The binary pulse-width lamps flashed once, indicating that the measurement was made. The pulse-width memory lamps in trunk 1 did not light, because no match pulse was generated.

324 The video pulser was pulsed again. This time, a match pulse was generated in trunk 1. The binary pulse-width lamps flashed once, indicating that the measurement was made. The pulse-width memory lamps in trunk 1 lit and remained lit, indicating the measurement and storage of pulse width.

325 The video pulser was pulsed again. The binary pulse-width lamps flashed, indicating the measurement of the incoming pulse width. Although a match pulse was generated, there was no further read-in to the trunk 1 pulse-width memories, because the memories had stored the pulse-width data.

(c) PRF

326 The trunk was reset, and the pulse width was turned off for clarity. The PRF counter oscillator was set at 2500 cps instead of 250 kc. This increased the automatic reset time from 0.1 second to 10 seconds for convenience of visual demonstration.

327 The video pulser was pulsed once. Frequency and direction data were stored in trunk 1 and were indicated by the lighted lamps. The trunk 1 read-in started the PRF counters. This action was indicated by the blinking PRF counter lamps.

328 The video pulser was pulsed for the second time. This generated a match pulse in the usual fashion but had no effect on the PRF counter.

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329 The video pulser was pulsed for the third time (within 10 seconds of the first time). This generated the second match pulse, which stopped the PRF counter. The PRF lamps remained lit to indicate the time interval between the first and third intercepts; PRF was twice the inverse of this period. The video pulser can be pushed any number of times thereafter without changing the trunk memories.

(d) MULTI-RADAR SIGNAL HANDLING

330 Pulse handling of all measured parameters of a single radar had now been demonstrated. The system is capable of handling many radars at the same time as long as the pulses are not time coincident. The signal handling in the presence of more than one radar was demonstrated next.

331 To demonstrate signal handling, the complete system was energized, except for the read-out, and input condition 5 was used. All six radars were set to a desired frequency, direction, pulse width, and PRF.

332 A simulated radar was pulsed three times to insert data in trunk 1. Additional pulses indicated that no further entries of that radar were made. Another radar was pulsed three times to show that it was stored in trunk 2. Trunks 1 and 2 were reset and pulsed again, each three times in any order to show that the system could handle interleaved pulses. This action also showed that an intercept is stored in a trunk with receipt of the first pulse and that pulse width and PRF are correctly measured and stored with the associated frequency and direction data.

333 Three other radars were pulsed three times each to fill the trunks. Partial trunks are available for trunks 3, 4, and 5. The frequency bits stored are 1, 2, 4, and 8. For clarity in the demonstration, radars were set to give binary numbers having unique combinations of these numbers.

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Channels 20 and 36, for example, give identical combinations of the binary numbers. Channels 32 and 80 give binary numbers containing none of the stored bits.

(e) READ-OUT

334 The tests described in paragraphs 330 through 333 showed that signals were properly processed for a single radar, for several signals separately, and for signals that were intermixed. At the end of these tests, the trunks were filled. The read-out was switched to slow speed, and each line of the trunk was read out in 5 to 10 seconds. During this period, lights remained on in both the read-out indicator row and the trunk row to permit verification of correct read-out.

335 The read-out then proceeded to sequence through the navigation trunks. This was visually indicated by (1) the NAV lamp lighting, and (2) the lamps next to each computer trunk going on and off in sequence. On the first navigation row, all the record track lamps lit, except the first (sync) track lamp. On the other nine navigation rows, none of the 15 data track lamps lit, because there was no navigation data to record. The first track lamp, however, remained on. After the read-out had sequenced through the navigation trunks, it passed on to the intercept data trunks. This action was indicated by the NAV lamp going out and the data lamp going on.

336 At the time of passing from trunk 1 to trunk 2, the read-out device reset trunk 1. This action was indicated by the reset lamp in trunk 1 flashing once and all the lamps in trunk 1 going out. In a similar manner, the read-out was allowed to sequence through, record, and reset the remaining trunks.

337 With the six radars set up to the desired frequencies and directions, the test-problem generator was made

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to pulse repetitively. This action refilled the trunks when they were read out. The read-out was placed on slow automatic cycling so that continuous storing and read-out of radars could be demonstrated.

338 The read-out rate was increased to 1 second per line of read-out information and permitted to operate for a few minutes. The read-out rate was set to full normal speed for a few minutes.

(3) IMPORTANT SIGNAL-PROCESSING DETAILS

(a) NOISE REJECTION

339 One noise-rejection feature is accomplished by requiring time coincidence in the frequency and direction analyzers. With input condition 5, the direction analyzer was switched on and the frequency analyzer was switched off. The video pulser was pulsed once. This action inserted a video pulse and simulated a noise pulse in the direction channel.

340 The direction-channel lamp flashed once. No other lamp flashed, because there was no frequency-channel output to start the clock. There was no further processing or storage for this pulse.

341 To show the effects of a noise pulse in a frequency channel, the frequency analyzer was switched on and the direction analyzer was switched off. The video pulser was pulsed once. The frequency-channel lamp flashed once. No lamps flashed in any direction channel. The output from the frequency channel started the clock. The frequency output blocking-oscillator lamps flashed, indicating encoding the frequency-channel data into binary form. The coincidence lamp in the computer did not flash because there was no directional signal. No other lamps, including the computer binary drivers, flashed. The system recovered and was ready for real radar pulses.

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342 The VIM provides additional protection against the unlikely occurrence of a noise pulse in frequency and direction in time coincidence. This action was demonstrated by switching on the frequency and direction analyzers. The PRF counter was switched to the slow speed.

343 The video pulser was pulsed once. This action inserted a noise pulse in the frequency and direction channels. Since the pulses in frequency and direction were in time coincidence, the system treated them as a radar pulse, and they were processed and stored in trunk 1. The PRF counter started and, in 10 seconds, automatically reset the trunk.

344 In the more unlikely case of two noise pulses in the same frequency and direction, the PRF counter will still reset the trunk 10 seconds after the first pulse.

345 The actual reset time of 0.1 second was demonstrated by switching the PRF counter to normal speed. The video pulser was pulsed to insert a coincident frequency and direction pulse into the system, which resulted in processing and storage in the normal manner. The computer memories appeared as flashes of light, because they were reset in 0.1 second. The preceding steps demonstrated the rejection of noise-generated pulses.

(b) RELIABILITY OF RECORDED INFORMATION

346 It is necessary to have safeguards to ensure that information recorded on the tape correctly represents the information stored in the computer trunks. The read-out of information is cyclic, whereas the storage of signals in the computer is random. For example, the PRF counter in one trunk could be counting while the read-out is transferring the count to the tape. Thus, the recorded information would be incorrect.

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- 347 The VIM ensures recording verified information. This action was demonstrated by turning the read-out and the PRF counter to slow and using input condition 5. A few seconds before the read-out was to read out trunk 1, the video pulser was pulsed. This action lit some of the frequency and direction memories and started the PRF counter. When the read-out cycled through trunk 1, none of the read-out record track lamps lit because the information was not verified. There was no trunk 1 reset pulse.
- 348 The pulser was pulsed twice. This stopped the PRF counter and turned on the VIM lamp. The next time the read-out cycled through trunk 1, the record track lamps lit, corresponding to the lighted lamps in the computer memories. As the read-out passed to trunk 2, the trunk 1 reset lamp flashed and all the trunk 1 memories were reset.
- 349 The ambel performs two basic functions. Strong signals into the frequency receiver and signals near the crossover of frequency or direction channels are capable of triggering more than a single channel. The ambel (1) selects the correct channel to respond, and (2) where the signal is near the crossover, generates an interpolated channel response, thus effectively doubling the resolution of the system.
- 350 The ambel and interpolation features were demonstrated by using input condition 4. For clarity, the computer was turned off during these tests.
- 351 To show the action of the frequency-analyzer ambel, the signal generator was set to repetitive operation and the frequency set to the middle of a channel. With a signal level just above triggering, only a single channel lamp lit and an even number was produced in the binary number corresponding to that channel number. As the frequency was shifted to the crossover between channels, two adjacent channel lamps lit and an odd binary number was generated

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- corresponding to an interpolate channel between the two. This action demonstrated the interpolate function when two channels responded. The signal-generator frequency was reset to the center of a channel and the signal strength increased until three channel lights lit. The ambel, under these conditions, chose the center channel, and the binary output lights indicated an even number for the center channel.
- 352 The interpolate channel is variable in width, depending upon signal strength. To demonstrate this limitation, the frequency was set to the edge of a channel. As the signal strength was increased, the adjacent channel started to respond and the binary number indicated the interpolate channel. As the signal strength was increased further, the adjacent channels on either side of the correct channel responded and the ambel chose the center channel. On the average, the interpolate channel width was 4 percent.
- 353 Similarly, the direction-analyzer ambel was demonstrated by shifting the phase of one-half the r-f signal (to simulate a direction change). The direction-channel lamps indicated one or two channels and the corresponding even or odd binary number.
- (c) EMPTY-TRUNK FINDING
- 354 Signals need not be stored in order in the trunks. The empty-trunk finder will choose the first empty trunk. To demonstrate this action, input condition 5 was used and all six radars on the test-problem generator were set up. Five radar intercepts were stored in the five trunks of the computer. When the sixth radar signal was pulsed, it was not stored. The read-out of one trunk was simulated by resetting one trunk. When the sixth radar was pulsed, it was stored in the empty trunk. The same experiment was performed by resetting two trunks and noting that the new signal was stored in the first empty trunk. For clarity, the

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pulse-width, PRF, and read-out circuits were not energized during these tests.

(d) TELEVISION-INTERFERENCE SUPPRESSION

355 In areas of high-power television signals, the high horizontal-line scan frequency (10 to 20 kc) and the video signals will tend to tie up the receivers and computers for a large percentage of the time. Thus, radar signals will be lost.

356 To demonstrate the effects of television interference, a Yagi antenna, tuned to the television frequencies, was coupled into the sub-band 2 portion of the feasibility-demonstration system. Many signals (television channels 2, 4, and 5) were stored in the computer. With a large signal input from signal condition 4, the television signals were stored at the expense of the simulated radar. When the television-interference suppressor was energized, the interfering television stations were suppressed, permitting only the simulated radar signal to enter the system.

357 The television interference due to foreign stations was demonstrated by pulsing the r-f signal generator at horizontal-line scan frequencies corresponding to foreign television. The television-interference suppressor operated properly in response to these signals.

(e) LEFT-RIGHT INDICATION

358 The left-right indicator was fed directly from the r-f signal generator. An r-f pad in the left line attenuated the left signal. The r-f pulser was pulsed once. The right lamp in the output blocking oscillator flashed, the right lamp in the computer binary driver flashed once, and the right lamp in trunk 1 lit and remained on. To show a left indication, the right and left r-f cables were interchanged. Trunk 1 was reset and the r-f pulser was pulsed once. The

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right lamp did not light, indicating that the signal was on the left side.

(4) QUANTITATIVE MEASUREMENTS

(a) FREQUENCY

359 Sub-band 5: Using input condition 3, the signal generator was tuned across sub-band 5 at different signal levels to show frequency accuracy and the variable width of the interpolate channel. This showed that, on an average, the channel width is 4 percent and the average frequency accuracy is ± 2 percent. This test also demonstrated the dynamic range of the equipment. For this test, the computer may be turned off.

360 Sub-band 2: Using input condition 4, the signal generator was tuned across sub-band 2 at different signal levels. These measurements showed the same accuracy as was demonstrated in sub-band 5. For this test, the computer may be turned off.

(b) DIRECTION

361 Sub-band 5: Using input condition 1, with signal generator at 750 Mc, the full-scale mock-up was rotated to demonstrate direction accuracy. Signal level was changed to indicate that channel width is independent of signal level. The measurements were repeated at 480 and 1000 Mc.

362 Sub-band 2: Using input condition 2, with the antennas for sub-band 2 mounted in a 1/8-scale model of the aircraft, the model was rotated to demonstrate direction accuracy. Because of low gain of the antennas for sub-band 2, no signal-level measurements of signal-level variation were made. The measurements were made at 480, 750, and 960 Mc.

363 Using input condition 4, the accuracy of the direction analyzer was demonstrated using calibrated cable lengths

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corresponding to simulated angles-of-arrival of 0, 12, 36, and 48 degrees. These tests served to demonstrate that instrumentation accuracy is independent of the antennas.

(c) PULSE-WIDTH

364 Using test condition 3, the pulse width of the modulated r-f signal was varied. The binary output was totaled and compared with the reading of a Berkeley 1-Mc counter and a calibrated oscilloscope.

(d) PRF

365 Using test condition 3, the PRF of the modulated r-f signal generator was varied and compared with the binary information stored in the computer trunk. A Hewlett-Packard audio oscillator, monitored by a Berkeley counter, was used as the variable-PRF source. Appendix B contains an accurate conversion chart relating binary indication vs PRF.

(e) RESULTS OF FEASIBILITY DEMONSTRATION

366 The demonstration of the techniques and capabilities proceeded in a satisfactory manner. The basic signal processing principles and the important signal processing details (paragraphs 339 through 345) were successfully demonstrated to WADC personnel and invited guests on 4 February 1958.

367 A full demonstration and detailed quantitative measurements were made under the direction of the WADC project engineer on 14 through 17 January 1958. These data are summarized in a set of curves as follows. Figure 121 shows the bearing error of the feasibility-demonstration system vs frequency in sub-band 2. Calibrated lengths of cable were used to simulate angles of arrival, and the test results show the instrumentation accuracy of the system exclusive of antennas. The largest bearing error measured was 7 degrees. Figure 122 shows the bearing error of the system vs bearing in sub-band 2. These measurements were made with a 1/8-scale

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model of the aircraft and antennas, using sub-band 5 instrumentation. With few exceptions, the accuracy is within the ± 12 degrees specified.

368 Figure 123 shows the bearing error of the system vs frequency in sub-band 5. These measurements were made exclusive of antennas and demonstrate the instrumentation accuracy. The bulk of the error is a result of phase mistrack of the TWT's used as r-f preamplifiers. The maximum error is less than ± 12 degrees. TWT's that are tracked to greater accuracy than those used in the tests have since become available. We believe that the new TWT's would considerably reduce the errors shown in Figure 123. Figure 124 shows the bearing error of the system vs bearing in sub-band 5. These measurements were made with the full-scale mocked-up section of the aircraft. Here again, the error was within the ± 12 degrees specified.

369 Figure 125 shows the frequency error vs frequency in sub-band 2. Data were taken at two power levels to demonstrate an average accuracy of 2 percent and a minimum accuracy of 4 percent. Figure 126 shows these same data for sub-band 5 with the same conclusions with respect to accuracy.

370 Figure 127 shows the pulse-width error vs pulse width for the pulse-width analyzer. Note that the error is within the specified accuracy limits. Figure 128 shows the PRF error vs PRF. These errors are also within the specified accuracy limits.

13. RELIABILITYa. PRELIMINARY STUDIES

371 The preliminary reliability studies were devoted mainly to investigation of the applications of the Poisson distribution to reliability prediction. A considerable body of literature exists detailing prediction methods associated

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with "reliability block diagrams." As previously noted (references 4, 5, and 6), basic failure-rate data on components are quite sparse. Further, the association of failure-rate variation with environmental stress is not clearly defined. In general, a power-law variation is assumed, varying between third and seventh power depending on the type of environmental stress.

372 Despite these limitations, by judicious use of existing data, prediction can be made. Further, derating schedules based on good engineering practice can be drawn up. In general, all components used in the breadboard design of the feasibility demonstration model were derated by a factor of 4:1 for all stresses, such as dissipation, voltage, etc. All circuit breadboards were required to operate satisfactorily from -58 to +71 C as a minimum requirement. No mechanical stress (shock, vibration) was applied, since these problems fall naturally into a product design phase.

373 Some effort (consistent with time schedule) was placed in requiring that individual circuits and subsystems "fail safe." The failure of any component cannot be permitted to jeopardize the entire system. The extension to redundant circuits as a means of reliability improvement is apparent.

b. PREDICTIONS

374 Reliability calculations were made based upon the best engineering estimates that are available of the failure rates of parts used (reference 7). In all cases, the failure-rate data were assigned to the best available components--that is, MIL-specified, sub-miniature, etc.--that would be used in the service test models.

375 The survival probabilities were calculated on the basis of a ten-hour mission. The probability that no failure

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will occur in a ten-hour mission is 88 percent. Table IV summarizes a few of the calculations that were made.

TABLE IV
PROBABILITY OF SURVIVAL OF AN/DLD-2 IN A TEN-HOUR MISSION

Description of Failure	Probability (percent)
No failure	88
Failure of one trunk	88.5
Failure in pulse width only	90.2
Failure in frequency channel only	91.6
Failure in pulse width or PRF only	93

376 Several direct conclusions can be reached, based upon the data given in Table IV. First, increased reliability can be obtained with use of "better" components, and further derating of components. However, considerations of parts failure rates indicate that the improvement factors are relatively small because of the large numbers of components in use. This improvement is mandatory for service test models. Second, increased reliability can be obtained by means of redundant circuits. This technique was not exploited in the feasibility-demonstration model. However, serious consideration must be given to redundancies in the service test models. Third, any circuit failure must not jeopardize the mission. All circuits must fail-safe. Fourth, all circuits to be included in a service test model must be subject to review by the reliability group. The circuits must be margin-tested and environment-tested. Marginal checking procedures must be devised for complete subsystems. All of these factors would be applied to the design of a service test model.

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14. ACKNOWLEDGEMENTS

377 The interest and cooperation of WADC contributed greatly to the successful demonstration of the breadboard model of the AN/DLD-2(XA-1). The bulk of information relative to reliability studies was obtained from staff members of RCA and ARINC.

D. CONCLUSIONS

378 The AN/DLD-2(XA-1) is capable of measuring and recording electromagnetic ferret reconnaissance data in the 30 to 1000 Mc range. The measurement capabilities are:

1. Frequency: The average accuracy is ± 2 percent with a maximum error of ± 4 percent at some signal levels.
2. Direction: The direction accuracy is ± 12 degrees over each 120-degree azimuth sector. This accuracy is independent of frequency or signal level. The measurement is made in the 60 to 1000 Mc range. Left-right indication only is provided in the 30 to 60 Mc range.
3. PRF: The PRF accuracy is ± 2 percent over the 20 cps to 10 kc range.
4. Pulse width: The pulse-width accuracy is ± 5 percent or 1 μ sec over the 1 to 100 μ sec range.
5. Dynamic range: The measured dynamic range is between 46 and 60 db as a function of frequency. This is less than the specified dynamic range. This limitation was due to the restricted dynamic range of commercially available distributed amplifiers. Properly designed distributed amplifiers would correct this limitation.
6. Television-interference suppression: It was demonstrated that with television signals and no television-interference suppression, the system is badly jammed. With the television-interference-suppressor circuitry, it was shown that television signals are excluded and that stronger pulsed radar signals are properly handled.

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379 The direction measurements performed in sub-band 2 were pessimistic in that they combined the most difficult antenna siting problems with the most difficult r-f instrumentation problems (scaled to sub-band 5). The specified accuracy was maintained over 96 degrees of azimuth coverage abeam of the aircraft. A large contribution to direction error is a result of TWT tracking error. We have been assured by the TWT manufacturers (Huggins Laboratories Inc.) that better tracking can be obtained.

380 The signal-processing and data-handling capabilities of the AN/DLD-2 were successfully demonstrated. The system is capable of (1) receiving and storing up to 10 intercepts per second, and (2) properly associating navigational data with intercept data and of storage of navigational data on the output magnetic tape.

381 On the basis of the feasibility-demonstration tests, we believe that the AN/DLD-2 is feasible.

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PART II
RECOMMENDATIONS

382 We recommend that service test models of the AN/DLD-2 be built. The feasibility of the techniques have been demonstrated, and a system suitable for production would have the following physical characteristics:

Weight	595 pounds
Size	33.0 cu ft
Power consumption	900 watts

383 We recommend that the antennas used in sub-bands 1 and 2 be improved. These antennas (cavity-backed slots) should have higher front-to-back ratios, and additional model study would provide better amplitude and phase tracking. Circularly polarized antennas would require fewer antennas per sub-band. Thus, we recommend that loaded spiral antennas be investigated for use in these bands. We believe that lumped-loading provides a promising approach to the problem, and one that is attractive with respect to size and weight factors. Further, loaded spirals offer some advantage in improving the limited gain and poor impedance match of cavity-backed slot antennas.

384 We recommend that singly probed lines and directional couplers be used. This approach has been described in paragraph 99. Since multiple probing requires about 25 db of probe decoupling, as opposed to about 10 db of decoupling for the singly probed technique, the advantage of this technique is obvious. We believe that the complexity of multi-coupling offers little deterrent to the use of this technique.

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385 We recommend that television-interference elimination switching be accomplished electronically. The problems associated with switched grid-line bias in distributed amplifiers appear very amenable to solution.

386 We recommend that other memory devices be investigated for use in the computer. Some of these include alternative junction-device memories (PNPN diodes), hook transistors, and unijunction transistors. These devices consume less power than transistor flip-flops. We believe that they will become available with better-controlled characteristics and be supplied by more than a single manufacturer.

387 We recommend that the width of the sub-bands be reduced to 80 percent of an octave. In the higher-frequency region of a sub-band, where the antenna spacing becomes approximately $\lambda/2$ and, for large angles of arrival, line-probing problems present considerable difficulty due to tracking requirements imposed upon r-f circuits. In the lower-frequency region of a sub-band, where the antenna spacing becomes approximately $\lambda/4$, the lower instrumentation slope (db per degree) reduces the resolution capability of the direction finder. We believe that the reduction of the sub-band width is warranted, though this reduction will increase the required number of sub-bands from five to six.

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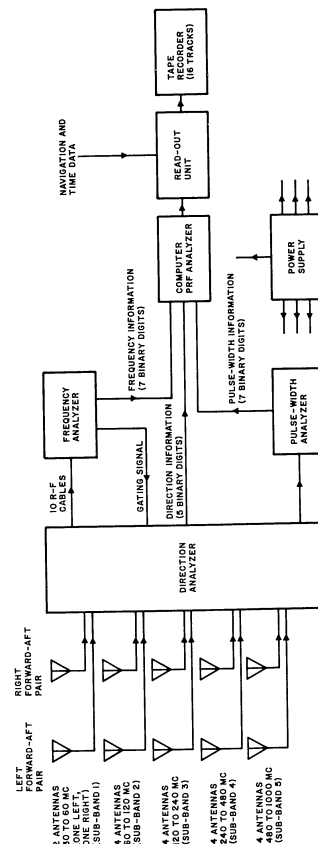


FIGURE 1. BLOCK DIAGRAM OF AN/DL-2(XA-1) SYSTEM

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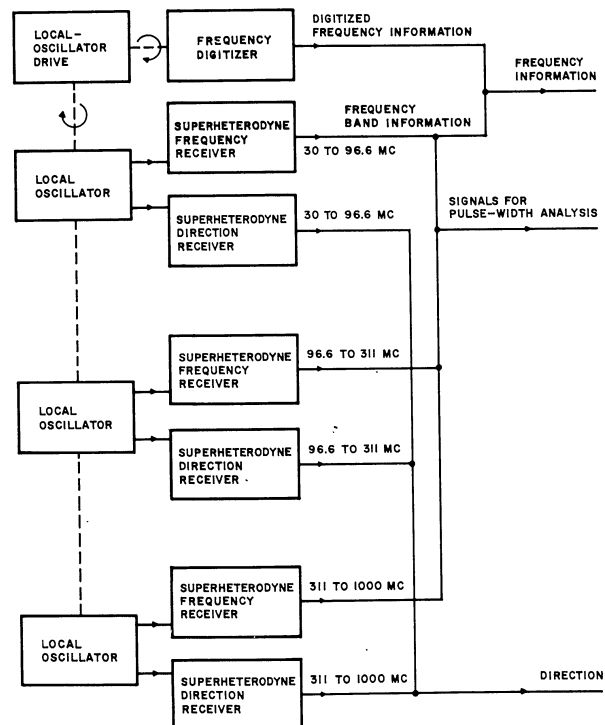


FIGURE 2. BLOCK DIAGRAM OF SCANNING SUPERHETERODYNE RECEIVER

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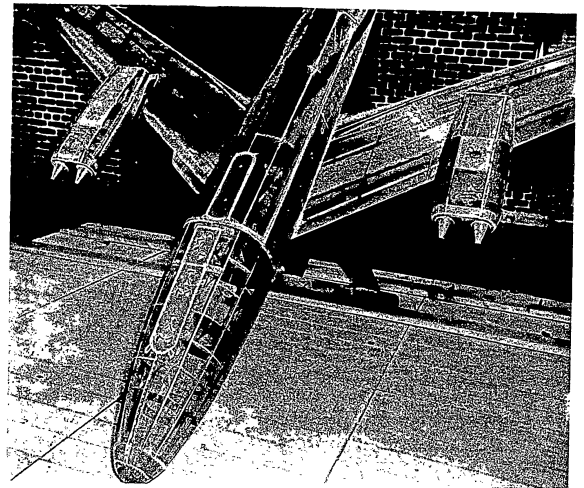


FIGURE 3. SCALE MODEL OF B-47, SHOWING SLOT ANTENNAS MOUNTED IN BELLY

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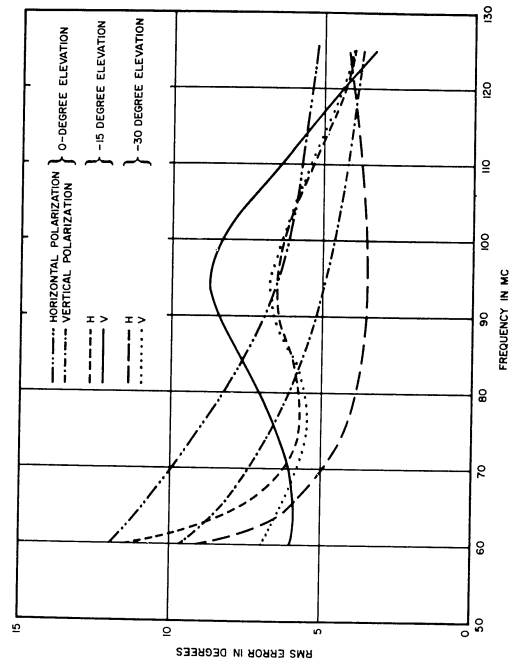


FIGURE 4. ANTENNA CONTRIBUTION TO RMS BEARING ERROR VS FREQUENCY IN SUB BAND 2

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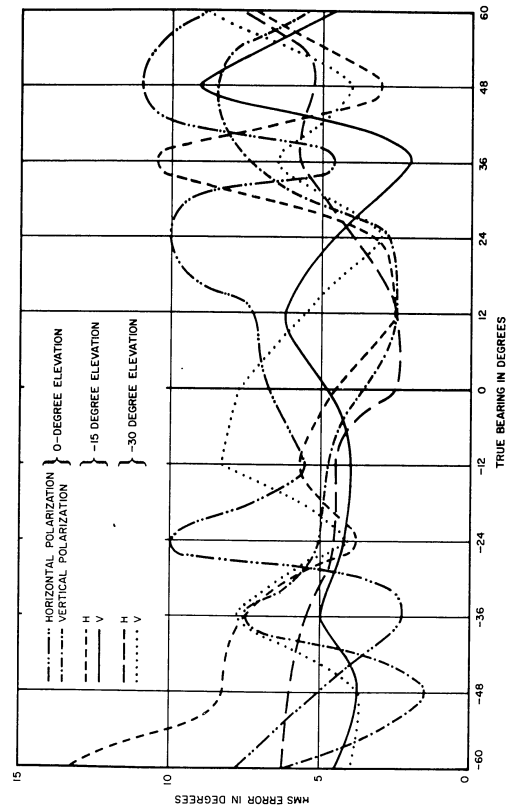


FIGURE 5. ANTENNA CONTRIBUTION TO RMS BEARING ERROR VS TRUE BEARING IN SUB-BAND 2

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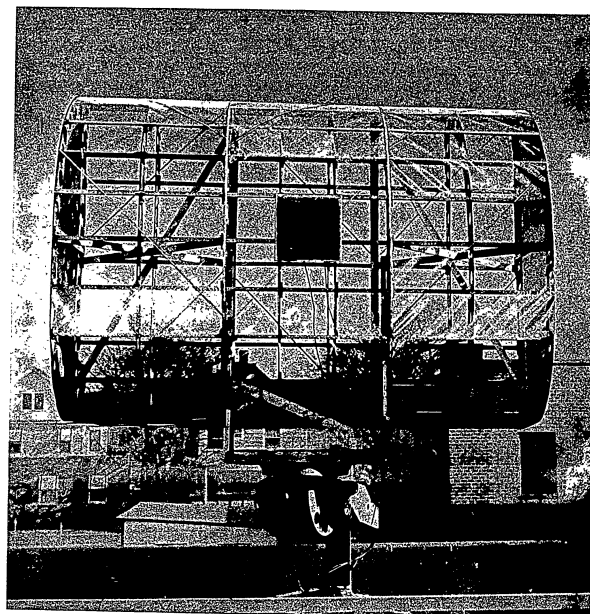


FIGURE 6. MOCK-UP OF SECTION OF B-47 AIRCRAFT, SHOWING SPIRAL ANTENNAS MOUNTED ON SIDE OF MOCK-UP

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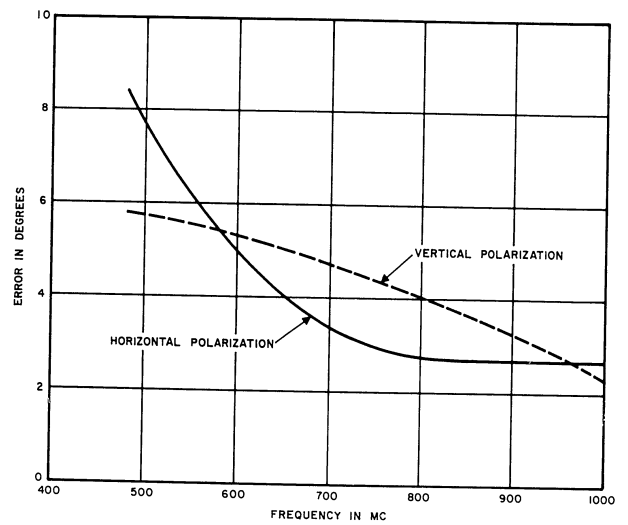


FIGURE 7. ANTENNA CONTRIBUTION TO RMS BEARING ERROR VS FREQUENCY IN SUB-BAND 5

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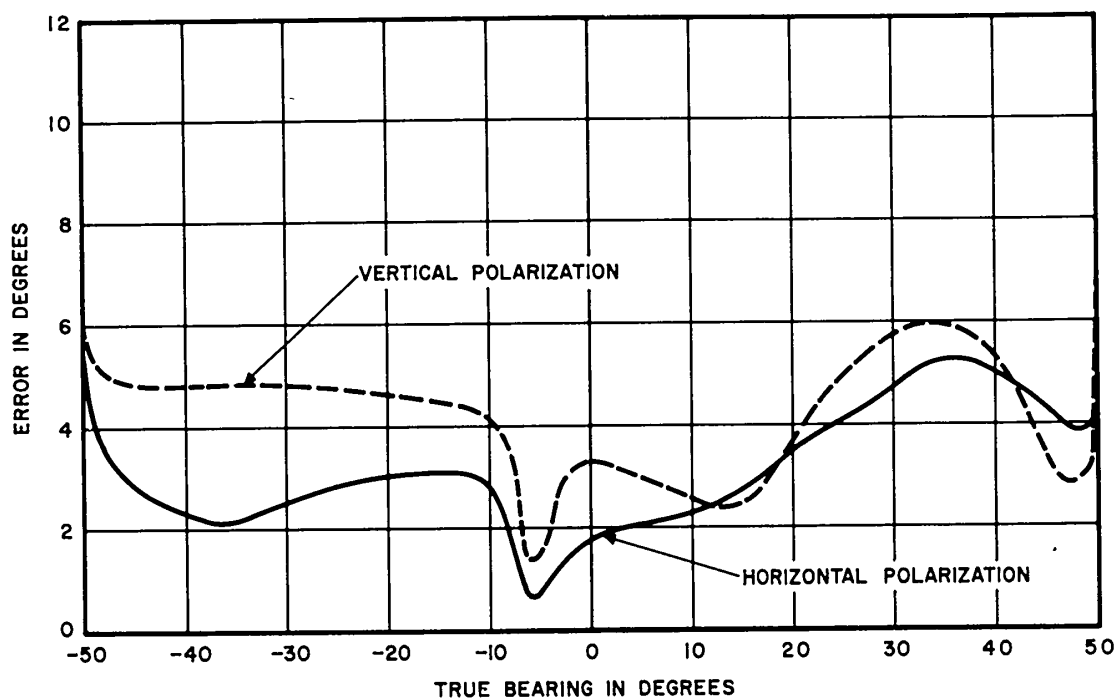


FIGURE 8. ANTENNA CONTRIBUTION TO RMS BEARING ERROR VS TRUE BEARING IN SUB-BAND 5

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FIGURE 9. ANTENNA RANGE

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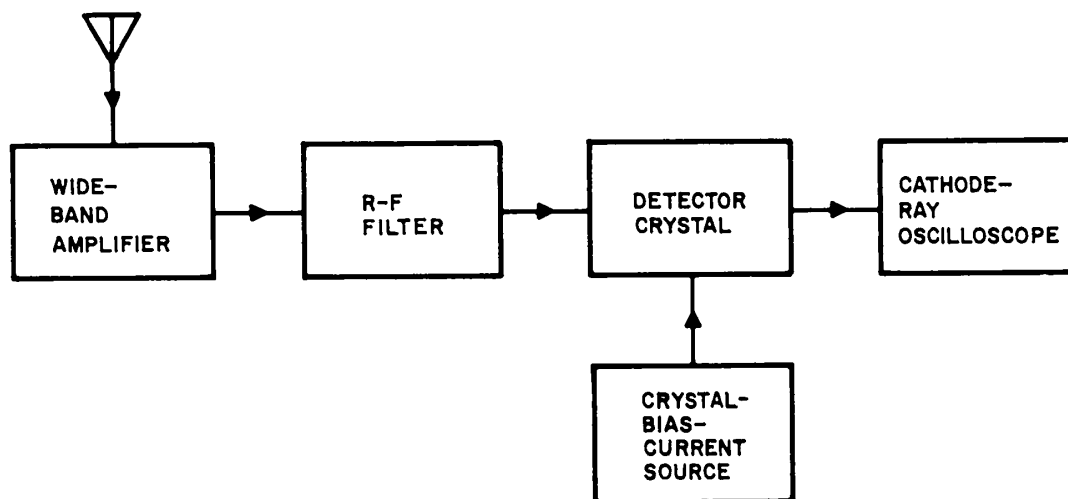


FIGURE 10. RECEIVING SYSTEM FOR TELEVISION-SIGNAL STUDY

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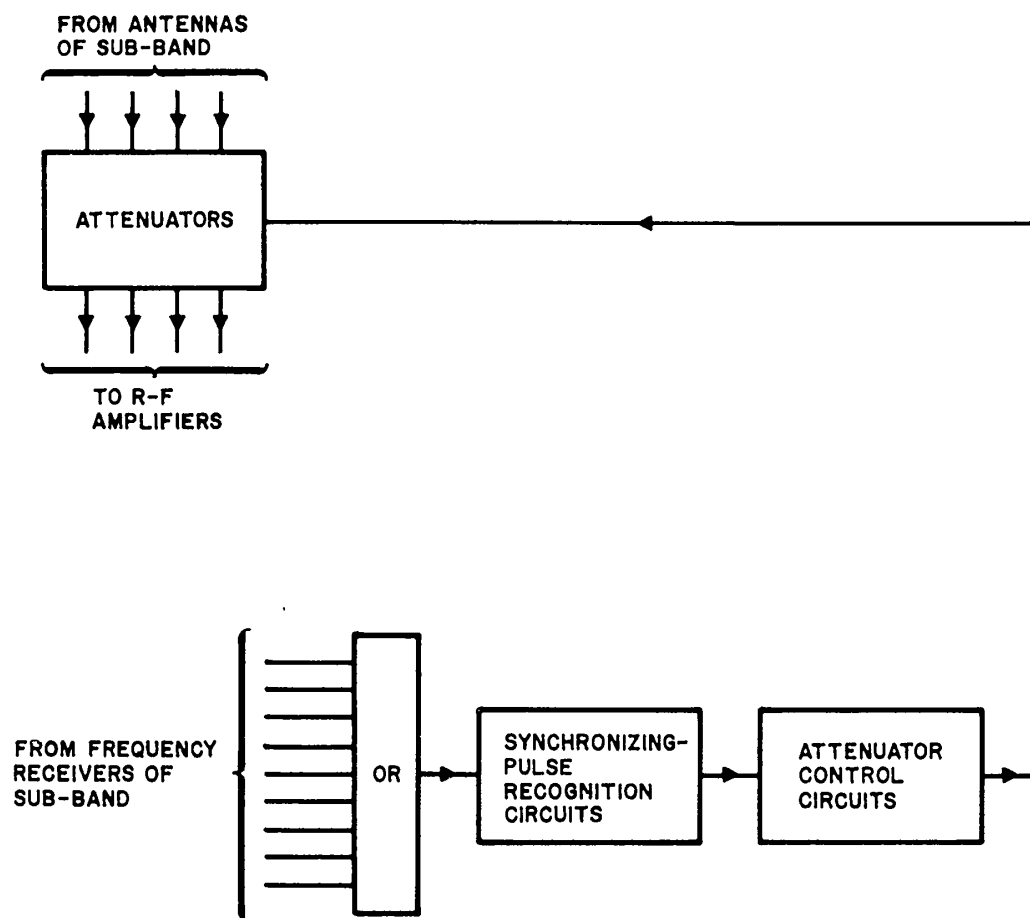


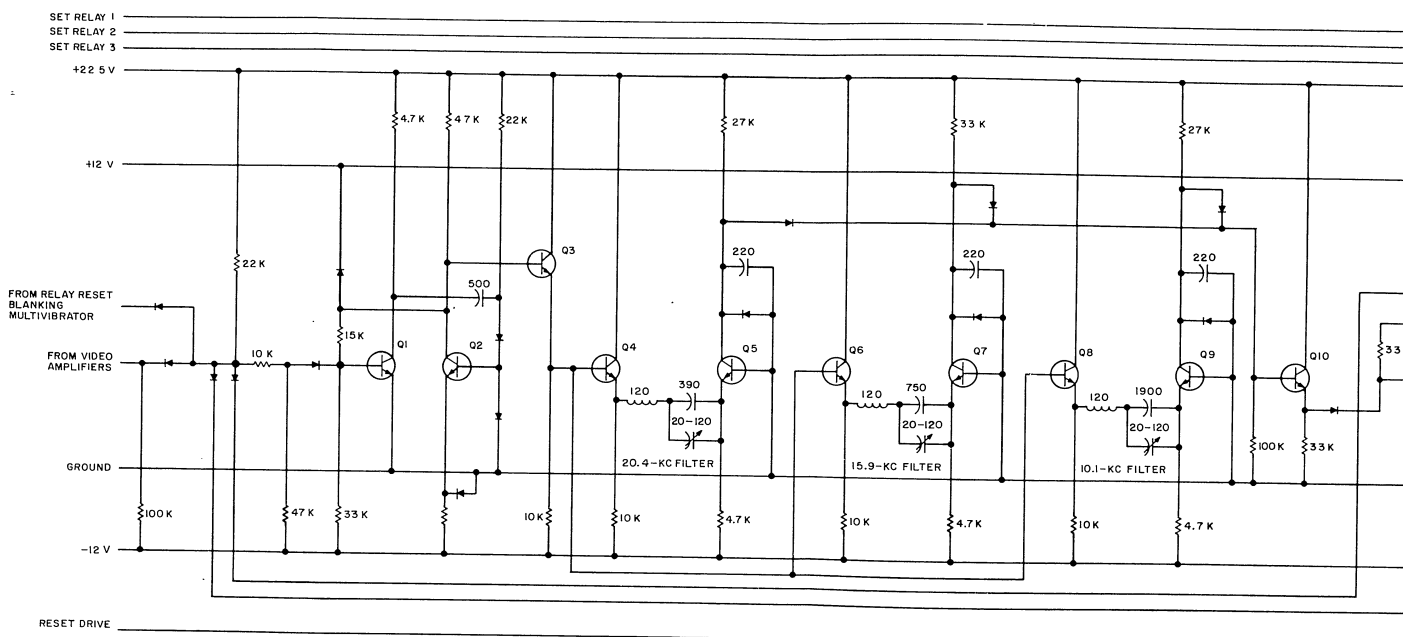
FIGURE II. BLOCK DIAGRAM OF TELEVISION-INTERFERENCE SUPPRESSOR

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UNLESS OTHERWISE NOTED:
 RESISTOR VALUES ARE GIVEN IN OHMS $\pm 5\%$, $\frac{1}{2}$ WATT
 CAPACITOR VALUES ARE GIVEN IN UUF
 INDUCTOR VALUES ARE GIVEN IN MH
 CRYSTALS ARE TYPE IN625
 TRANSISTORS ARE TYPE 2N118

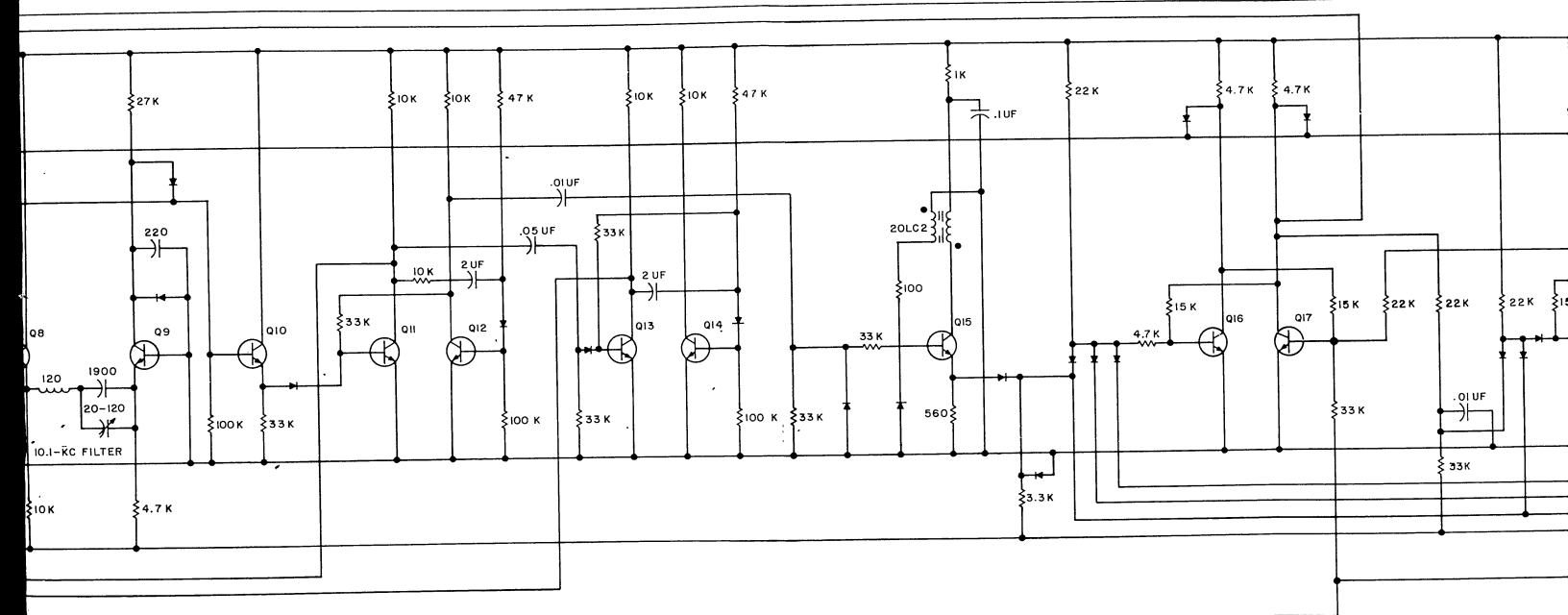
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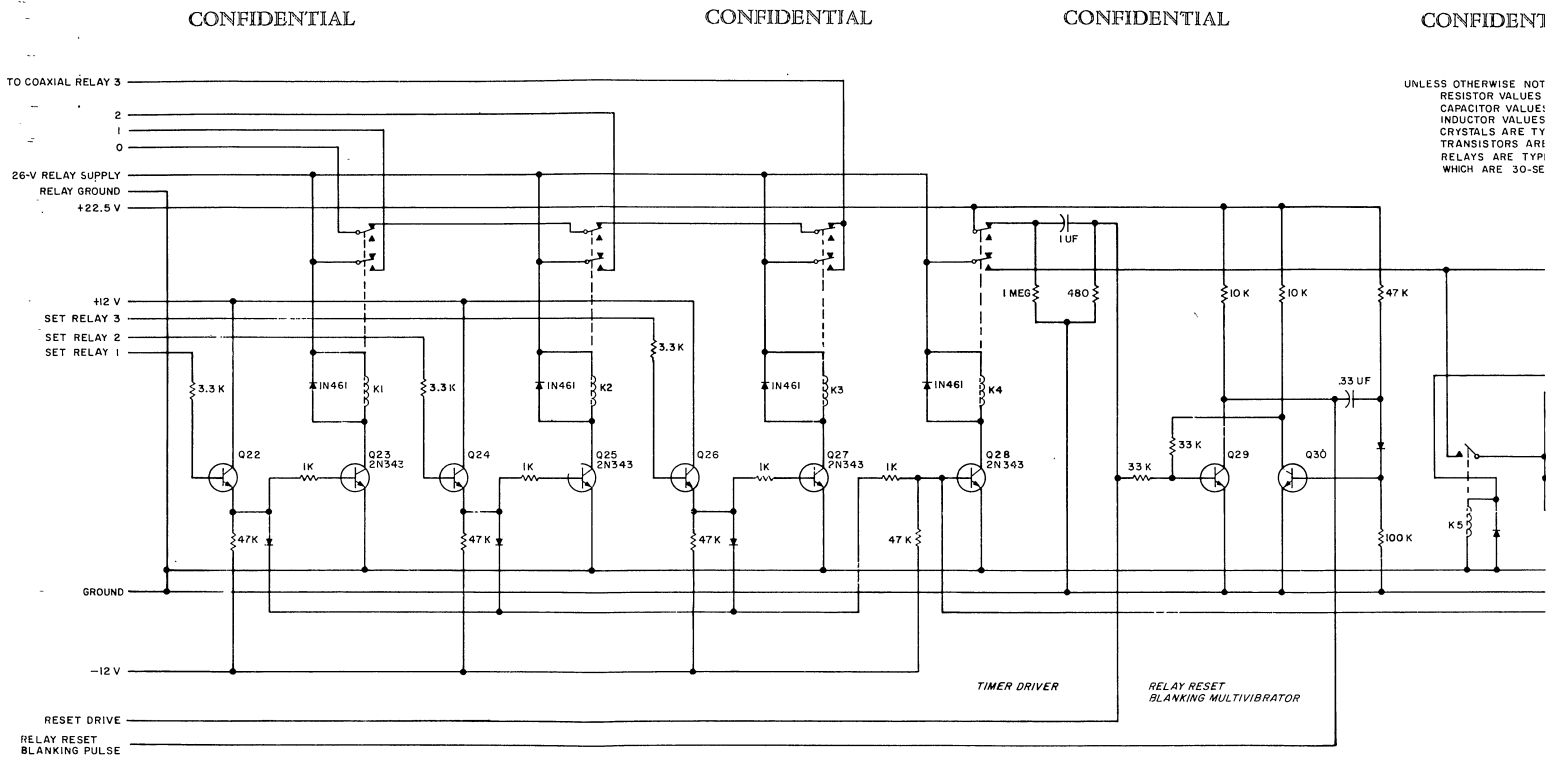


FIGURE 13. SCHEMATIC DIAGRAM OF ATTENTION

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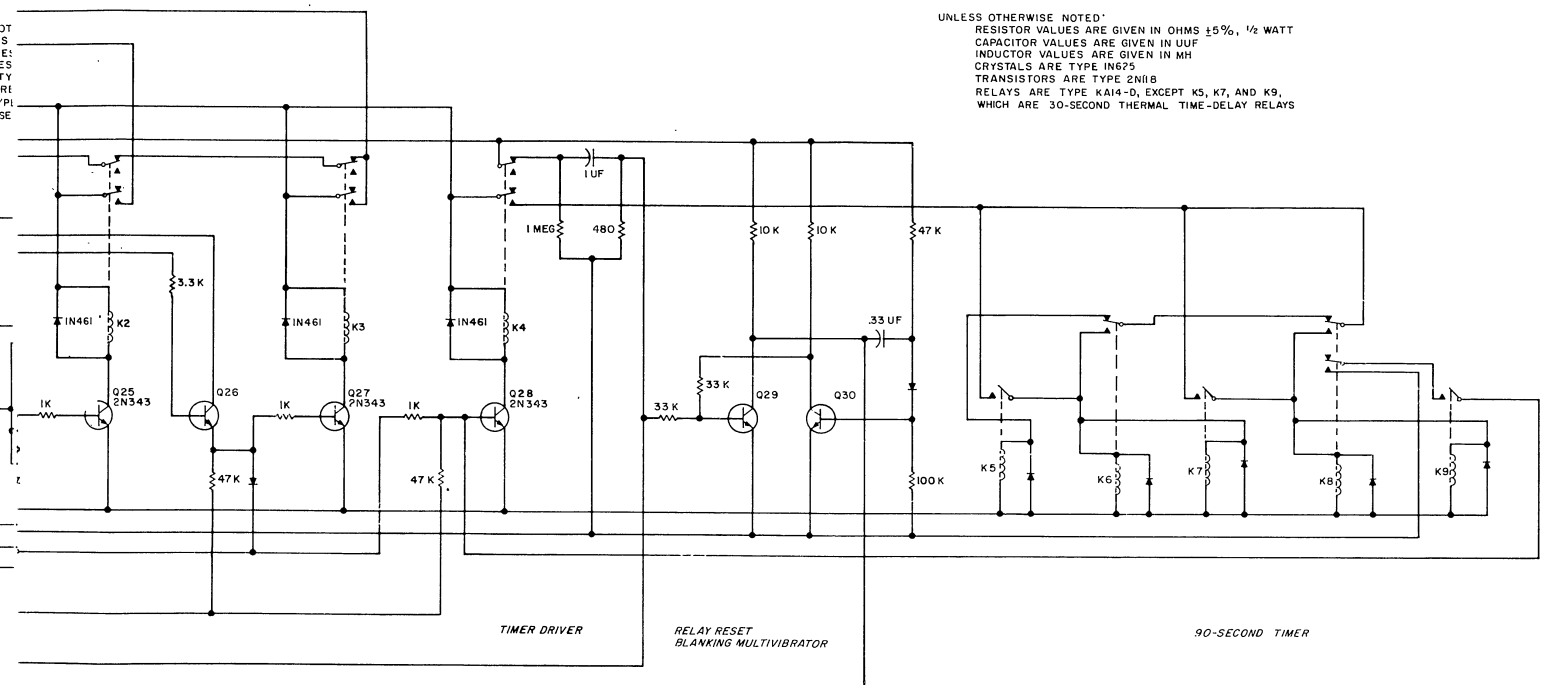


FIGURE 13. SCHEMATIC DIAGRAM OF ATTENUATOR CONTROL IN TELEVISION-INTERFERENCE SUPPRESSOR CIRCUITS

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FIGURE 13

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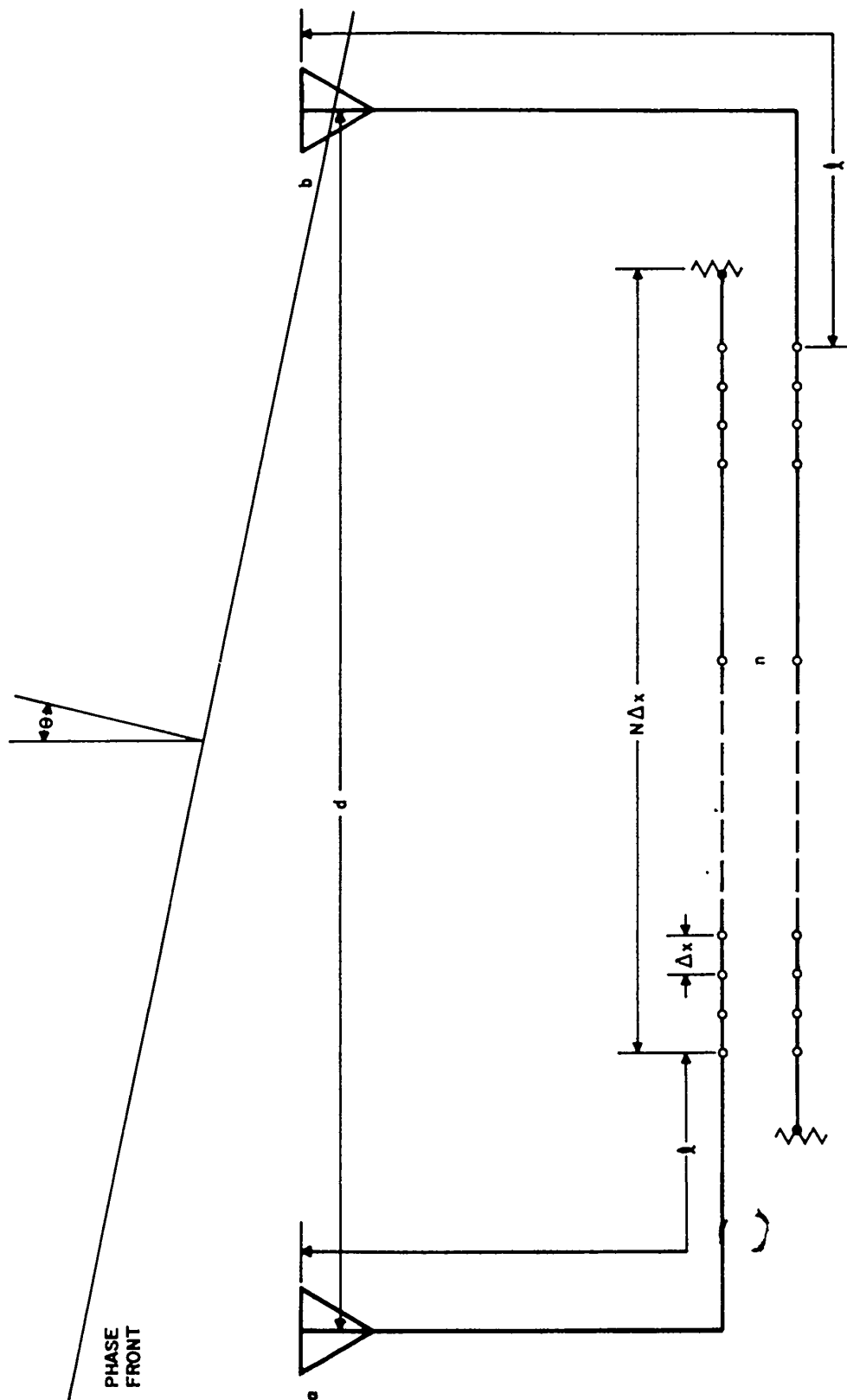


FIGURE 14. GEOMETRICAL CONSIDERATIONS IN PHASE-COMPARISON DIRECTION FINDER

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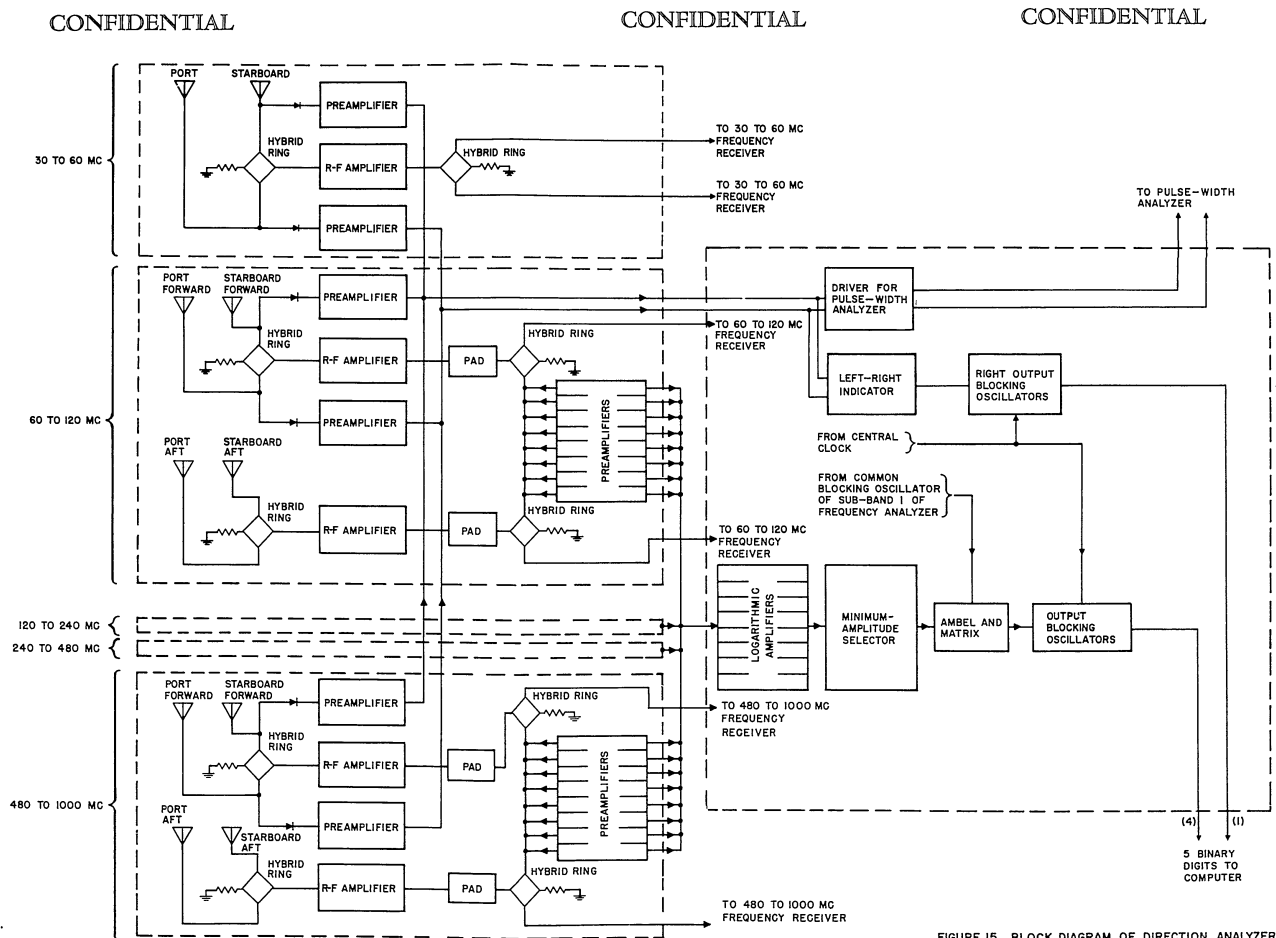


FIGURE 15. BLOCK DIAGRAM OF DIRECTION ANALYZER

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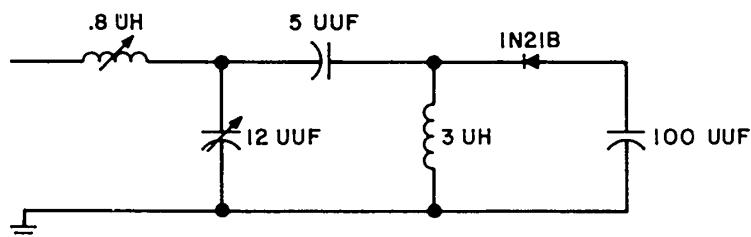
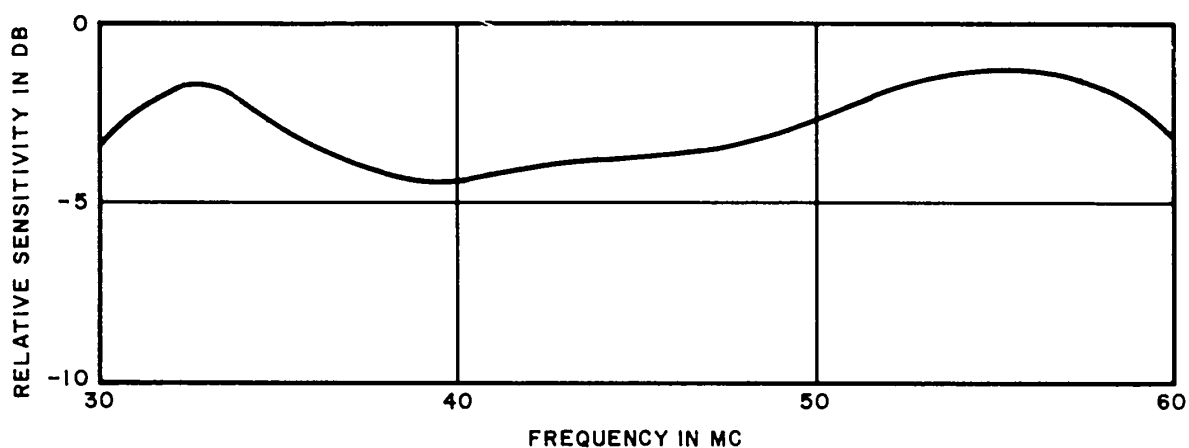


FIGURE 16. SENSITIVITY VS FREQUENCY OF DETECTOR MOUNT FOR SUB-BAND I
(WITH RESPECT TO TUNED DETECTOR MOUNT)

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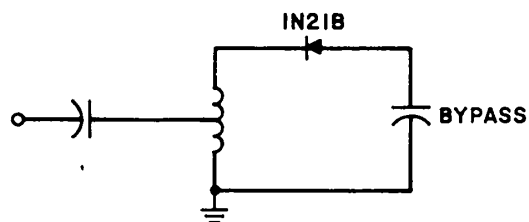
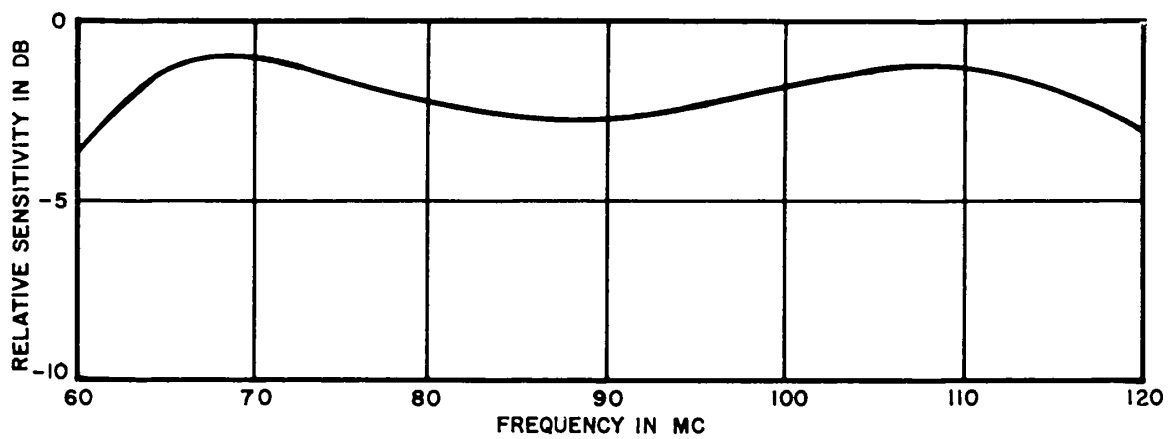


FIGURE 17. SENSITIVITY VS FREQUENCY OF DETECTOR MOUNT FOR SUB-BAND 2
(WITH RESPECT TO TUNED DETECTOR MOUNT)

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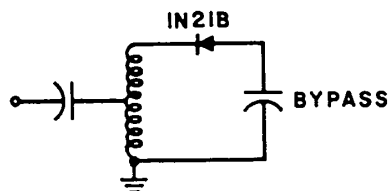
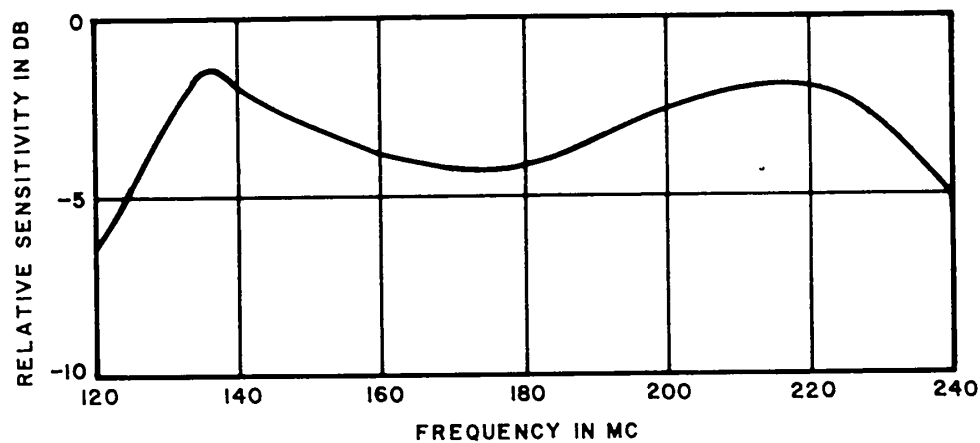


FIGURE 18. SENSITIVITY VS FREQUENCY OF DETECTOR MOUNT FOR SUB-BAND 3
(WITH RESPECT TO TUNED DETECTOR MOUNT)

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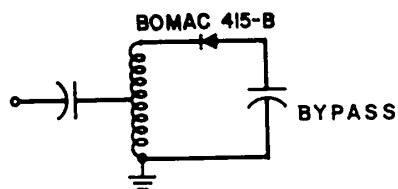
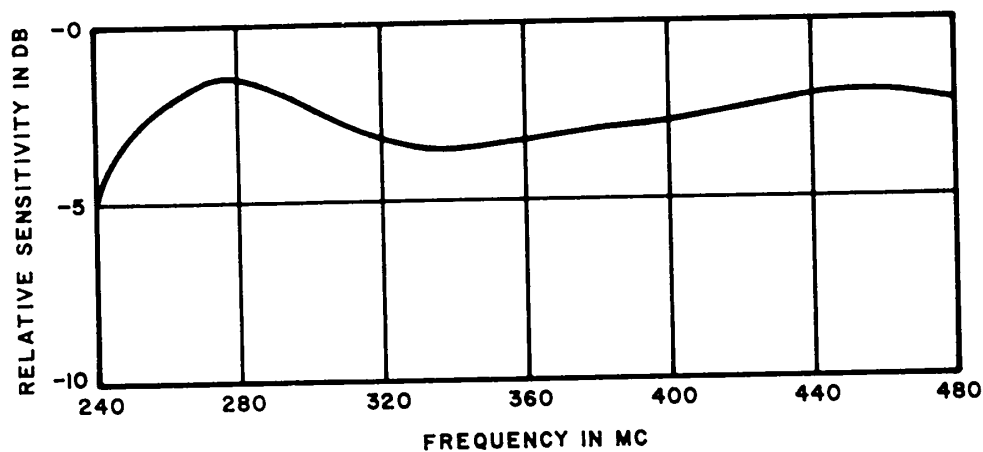


FIGURE 19. SENSITIVITY VS FREQUENCY OF DETECTOR MOUNT FOR SUB-BAND 4
(WITH RESPECT TO TUNED DETECTOR MOUNT)

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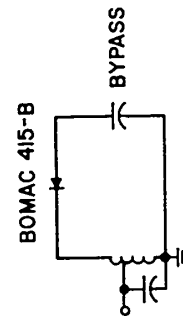
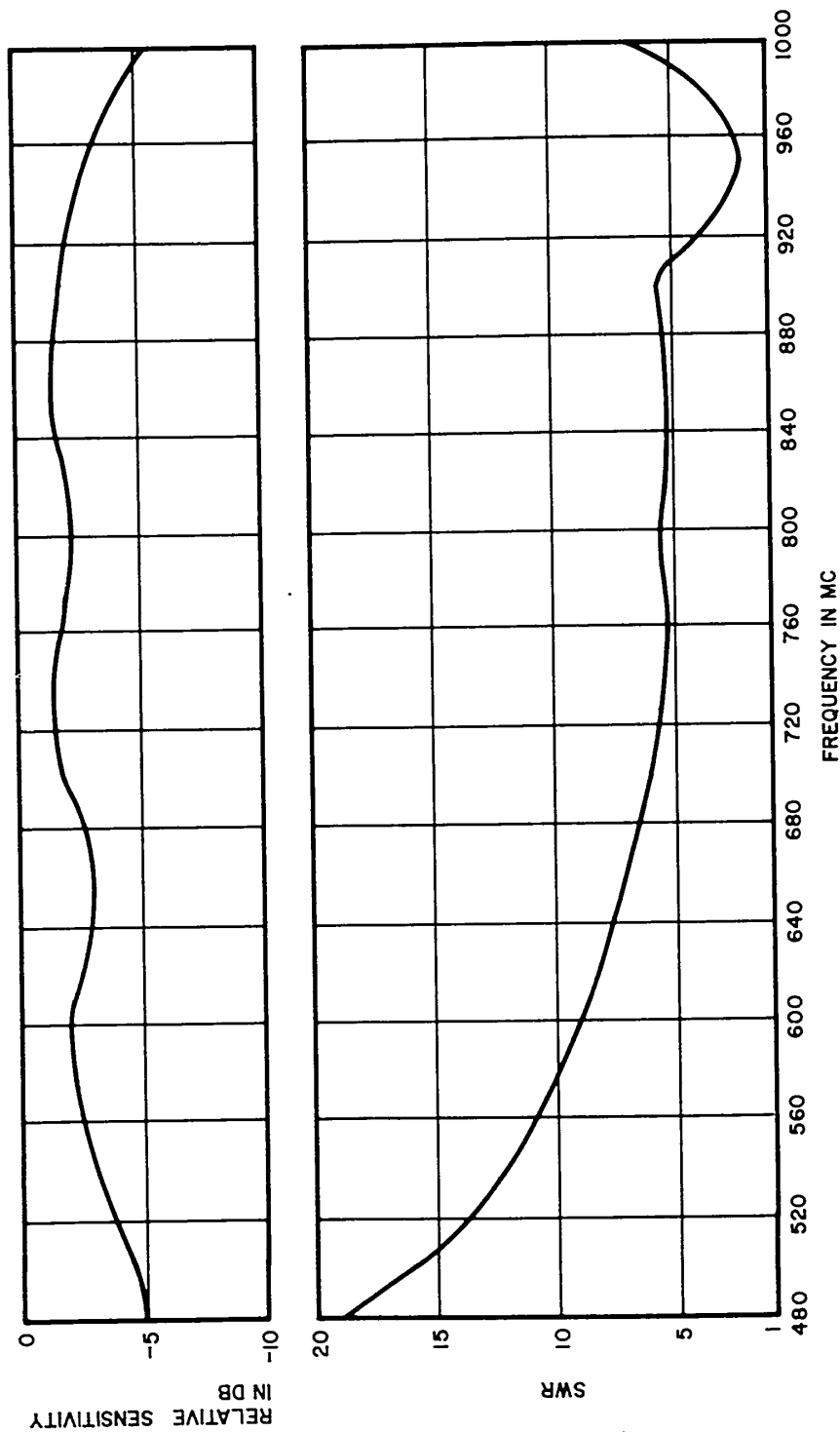


FIGURE 20. SENSITIVITY (WITH RESPECT TO TUNED DETECTOR MOUNT) AND SWR VS FREQUENCY OF DETECTOR MOUNT FOR SUB-BAND 5

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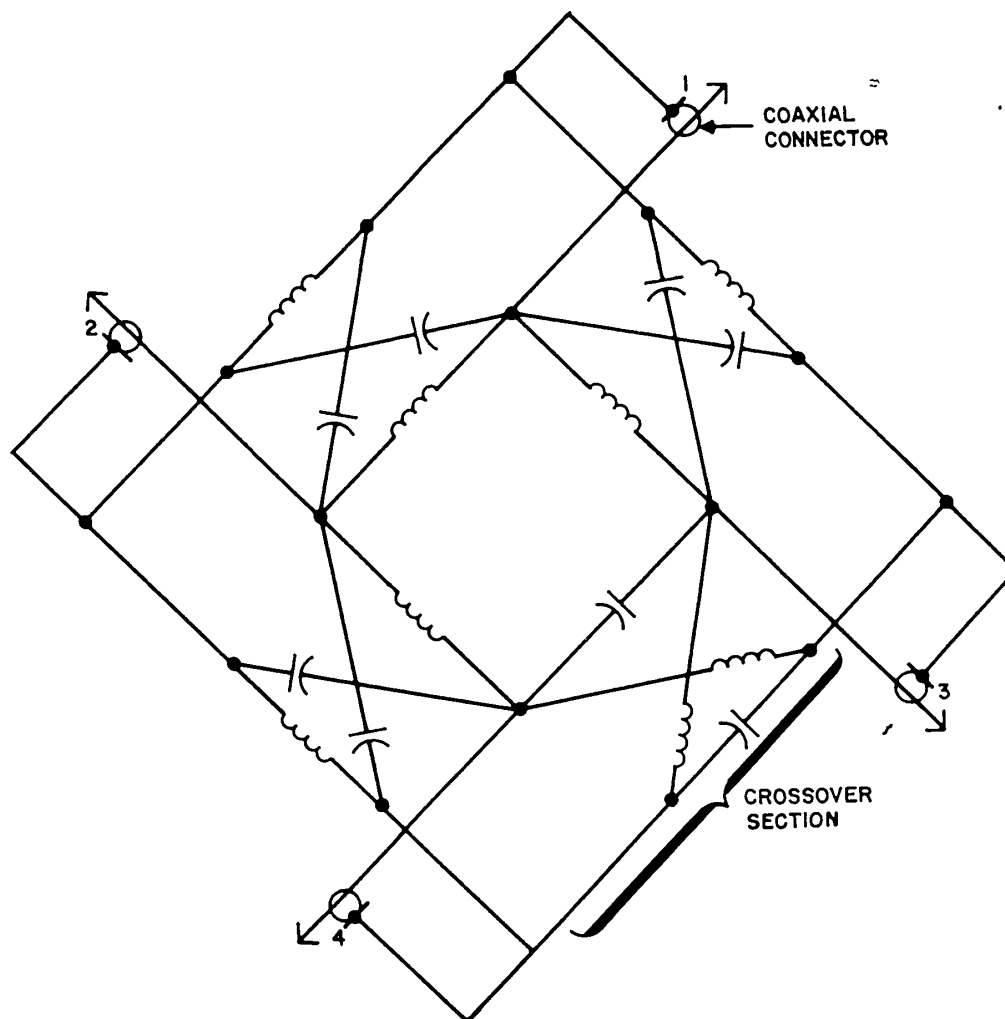


FIGURE 21. SCHEMATIC DIAGRAM OF LUMPED—CONSTANT HYBRID RING

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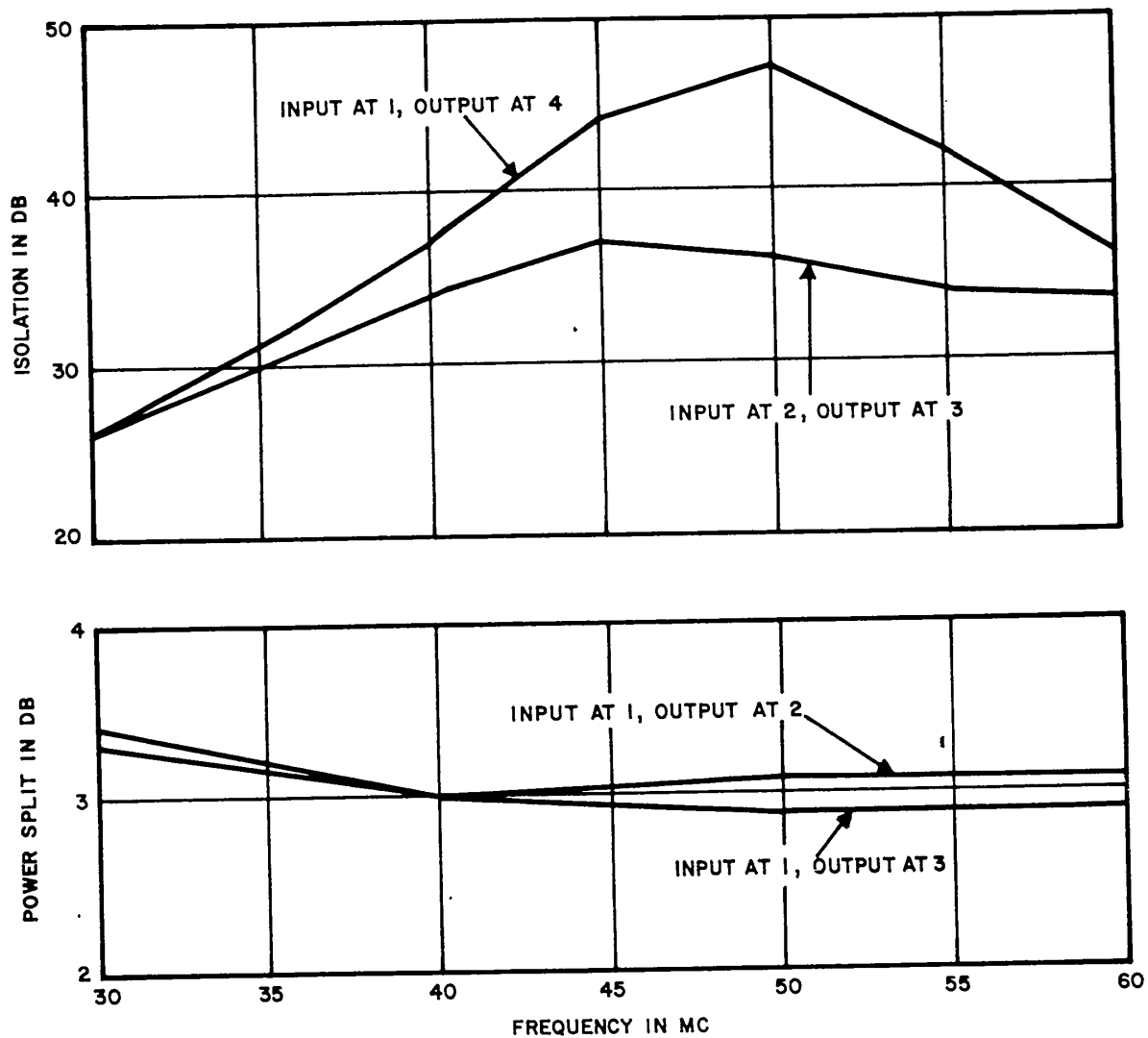


FIGURE 22. ISOLATION AND POWER SPLIT VS FREQUENCY OF HYBRID RING FOR SUB-BAND 1

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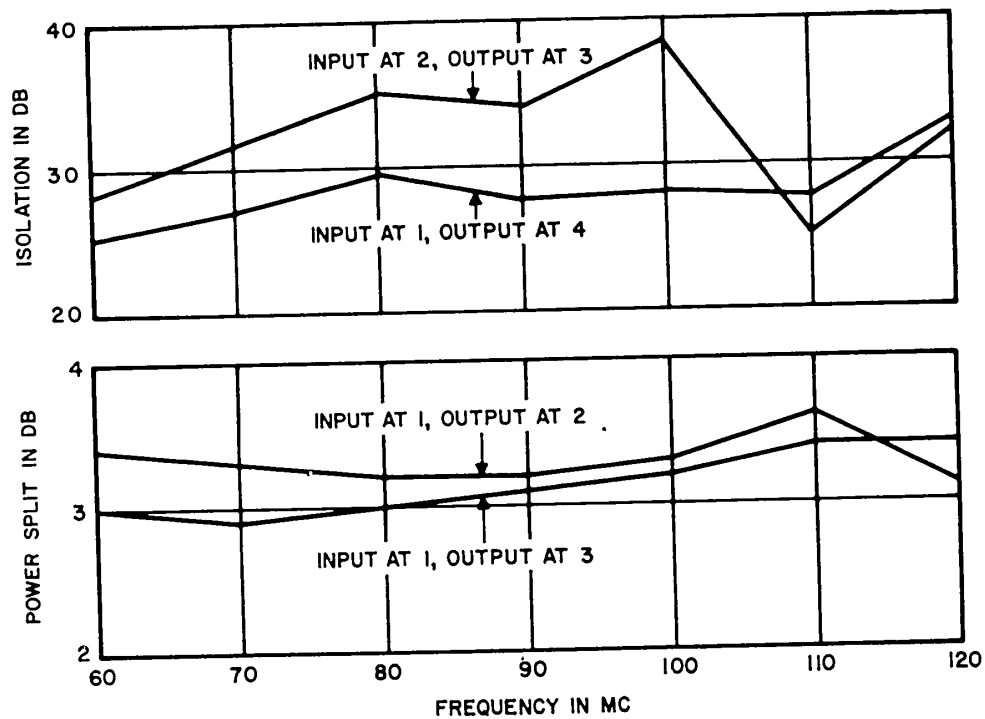


FIGURE 23. ISOLATION AND POWER SPLIT VS FREQUENCY OF HYBRID RING FOR SUB-BAND 2

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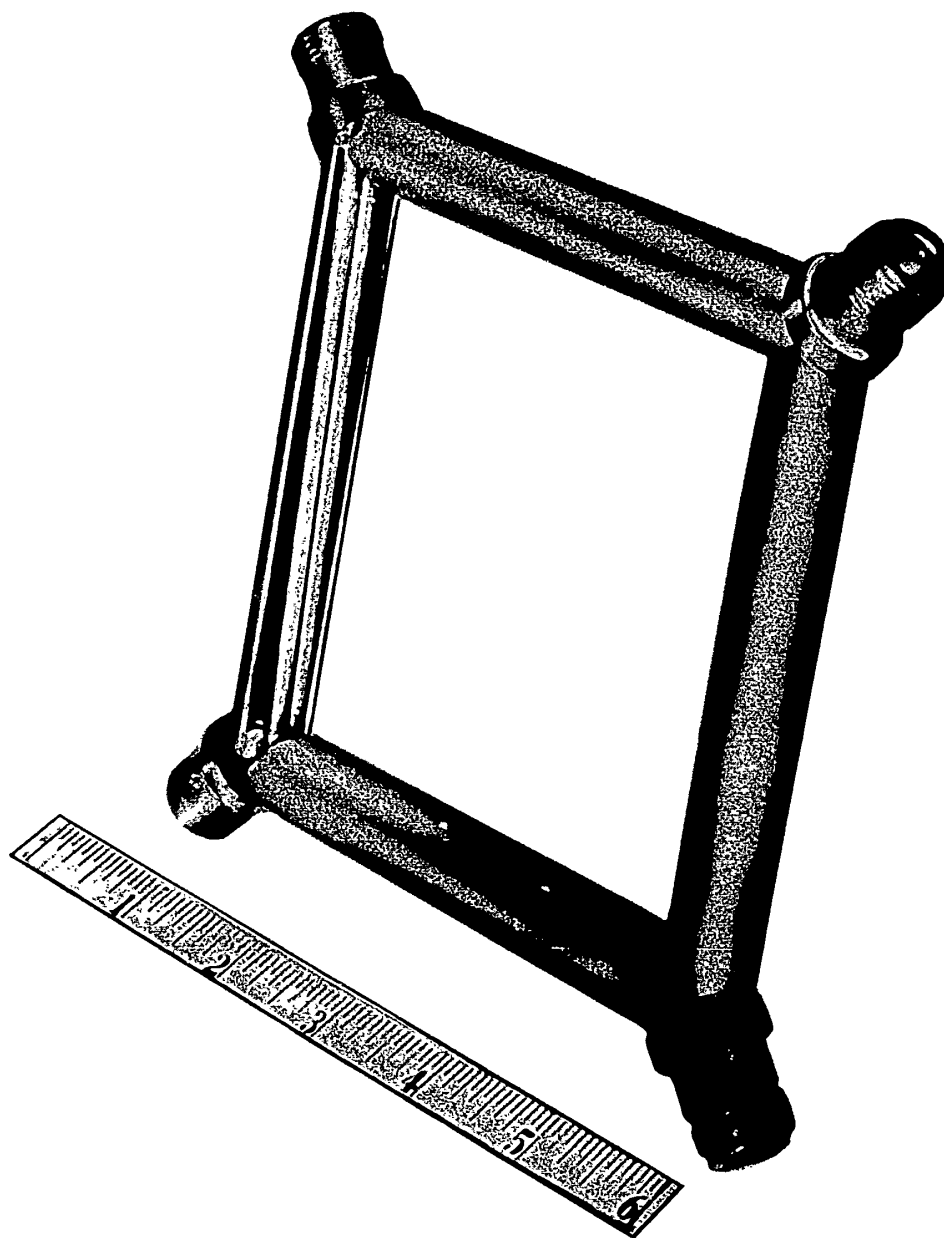


FIGURE 24. COAXIAL HYBRID RING FOR SUB-BAND 5

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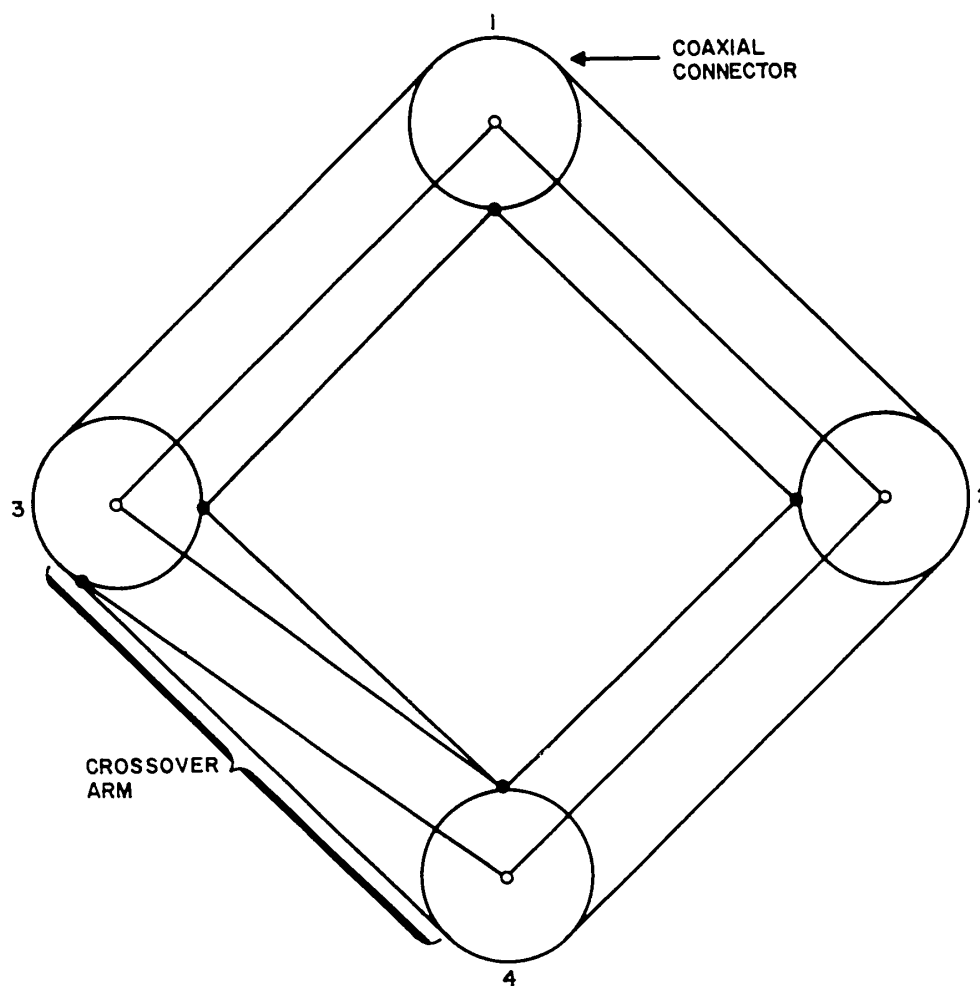


FIGURE 25. SCHEMATIC DIAGRAM OF COAXIAL HYBRID RING

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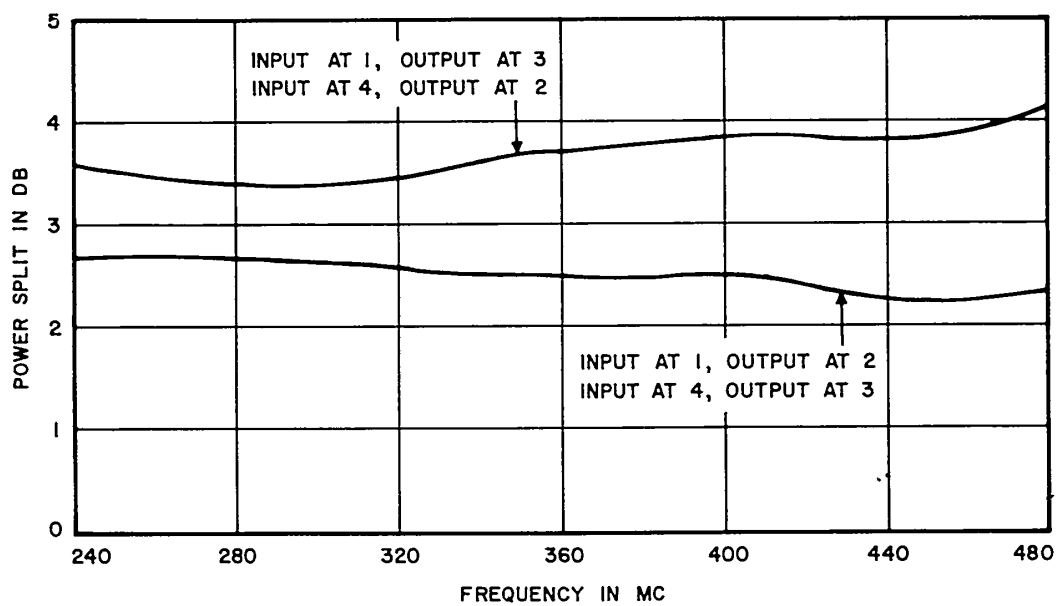


FIGURE 26. POWER SPLIT VS FREQUENCY OF HYBRID RING FOR SUB-BAND 4

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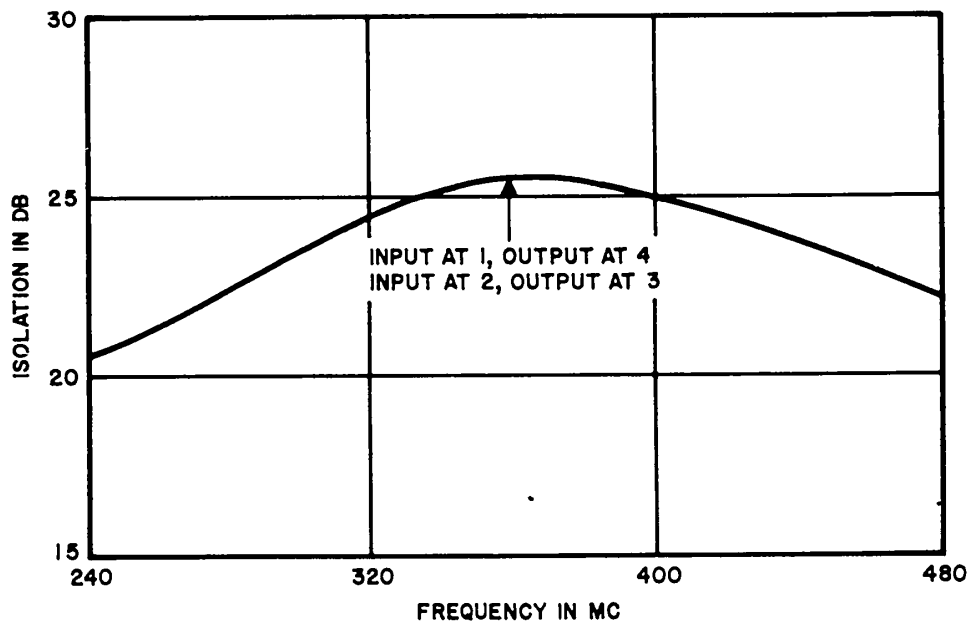


FIGURE 27. ISOLATION VS FREQUENCY OF HYBRID RING FOR SUB-BAND 4

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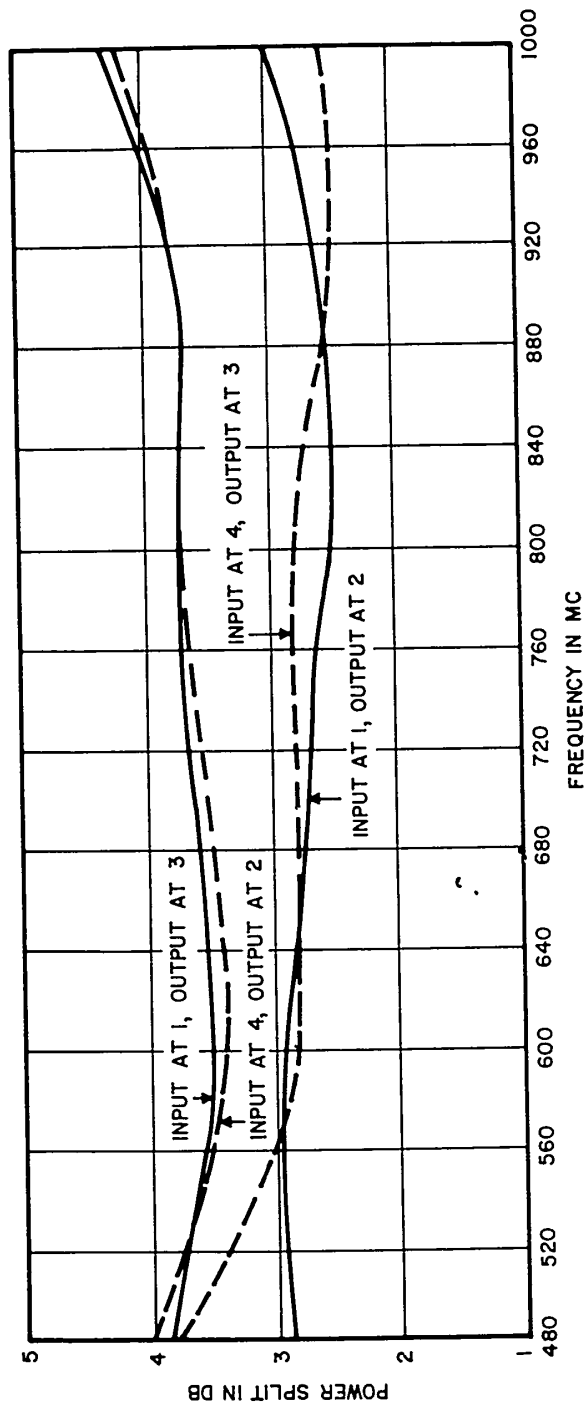


FIGURE 28. POWER SPLIT VS FREQUENCY OF HYBRID RING FOR SUB-BAND 5

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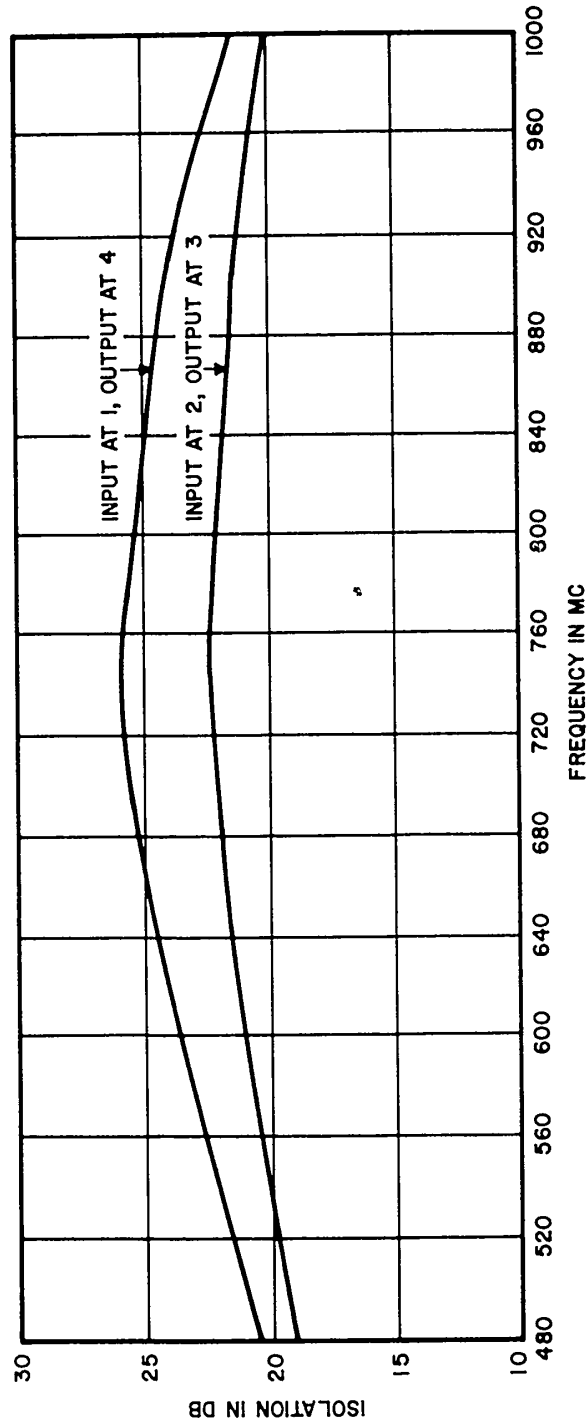


FIGURE 29. ISOLATION VS FREQUENCY OF HYBRID RING FOR SUB-BAND 5

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NOTE:
 $L_1 = L_2$ AND THEY ARE INDUCTIVELY COUPLED.
 PORTS 1, 2, 3, AND 4 ARE BNC CONNECTORS.

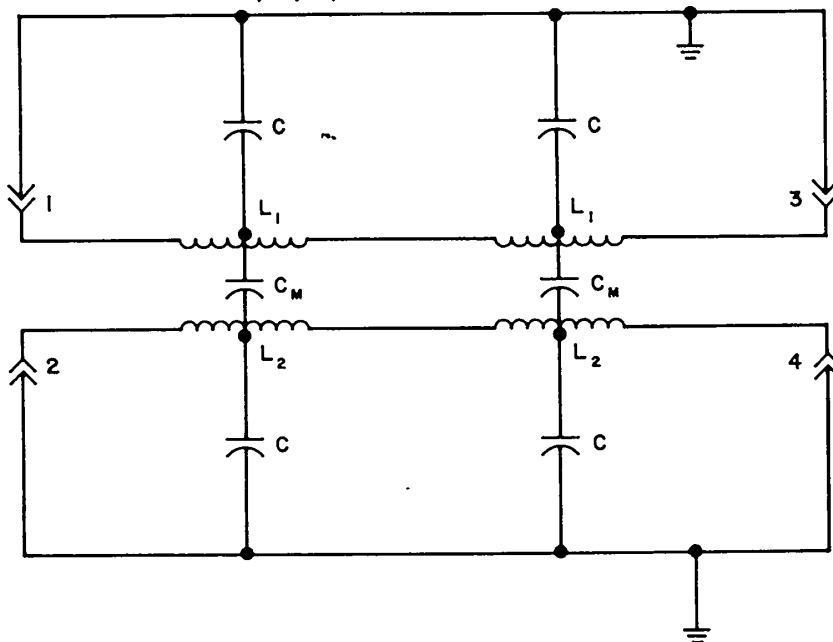


FIGURE 30. SCHEMATIC DIAGRAM OF LUMPED-CONSTANT DIRECTIONAL COUPLER

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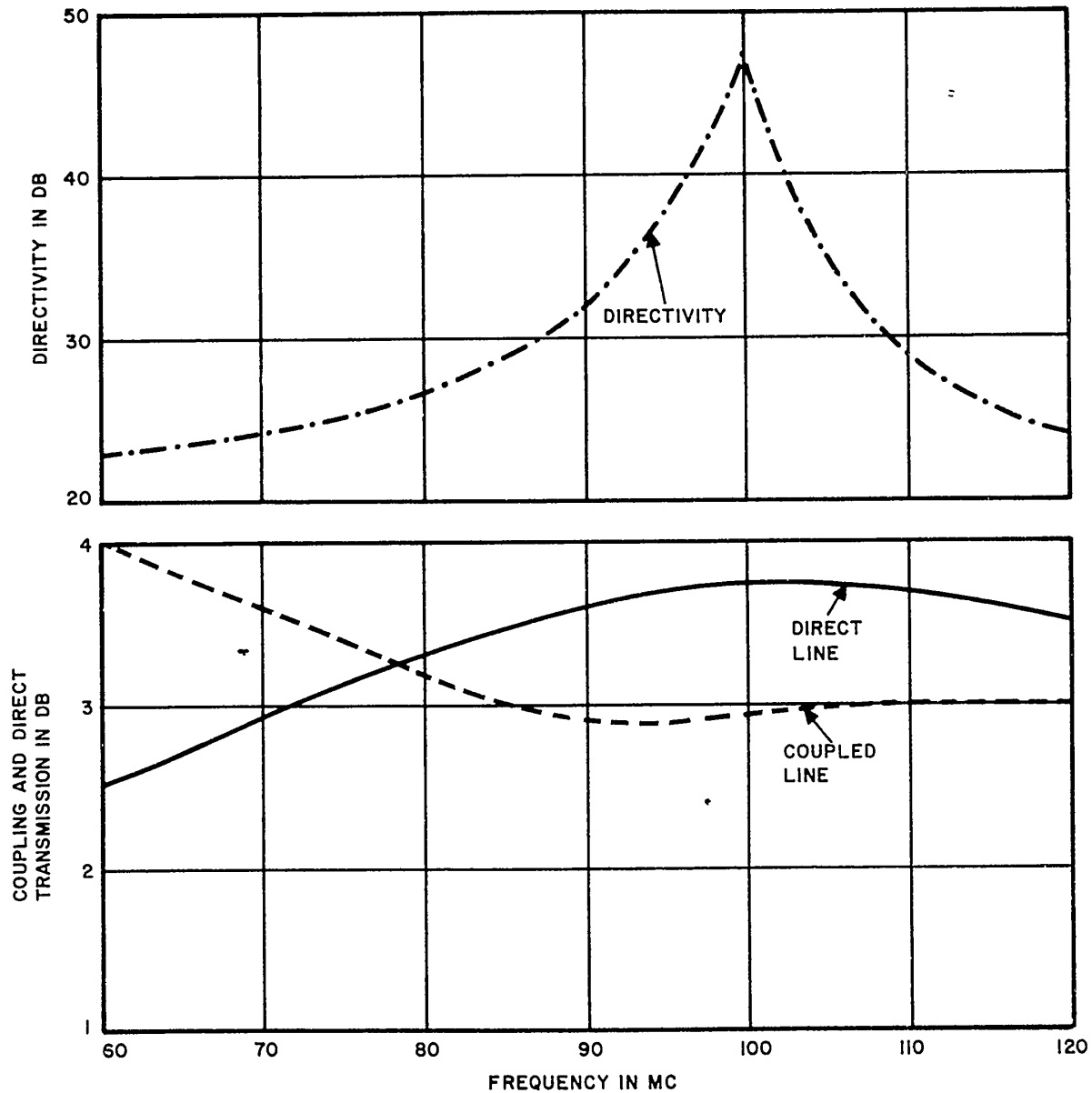


FIGURE 31. COUPLING, DIRECT TRANSMISSION, AND DIRECTIVITY VS FREQUENCY OF 3.0-DB DIRECTIONAL COUPLER FOR SUB-BAND 2

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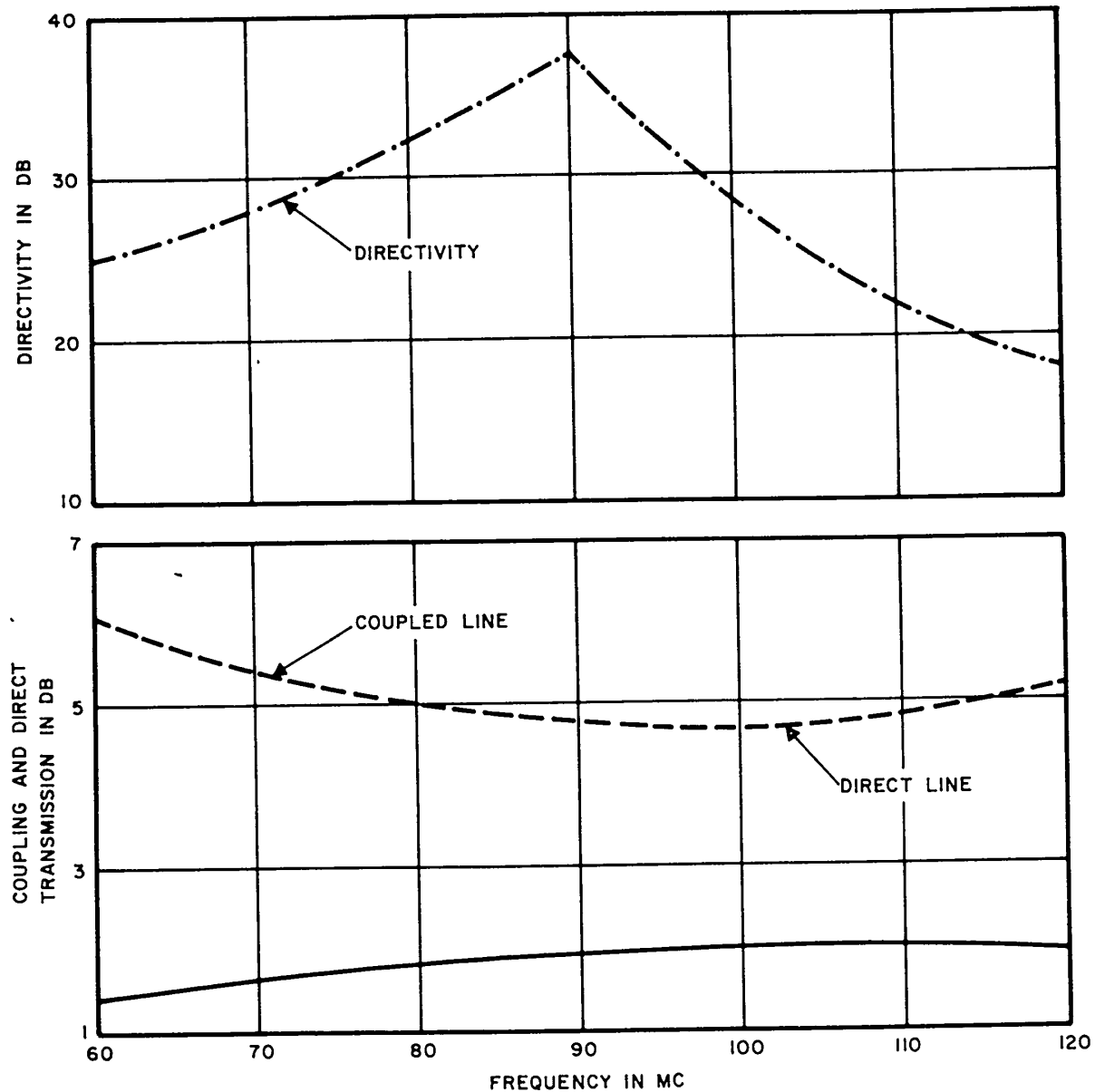


FIGURE 32. COUPLING, DIRECT TRANSMISSION, AND DIRECTIVITY VS FREQUENCY OF 4.8-DB DIRECTIONAL COUPLER FOR SUB-BAND 2

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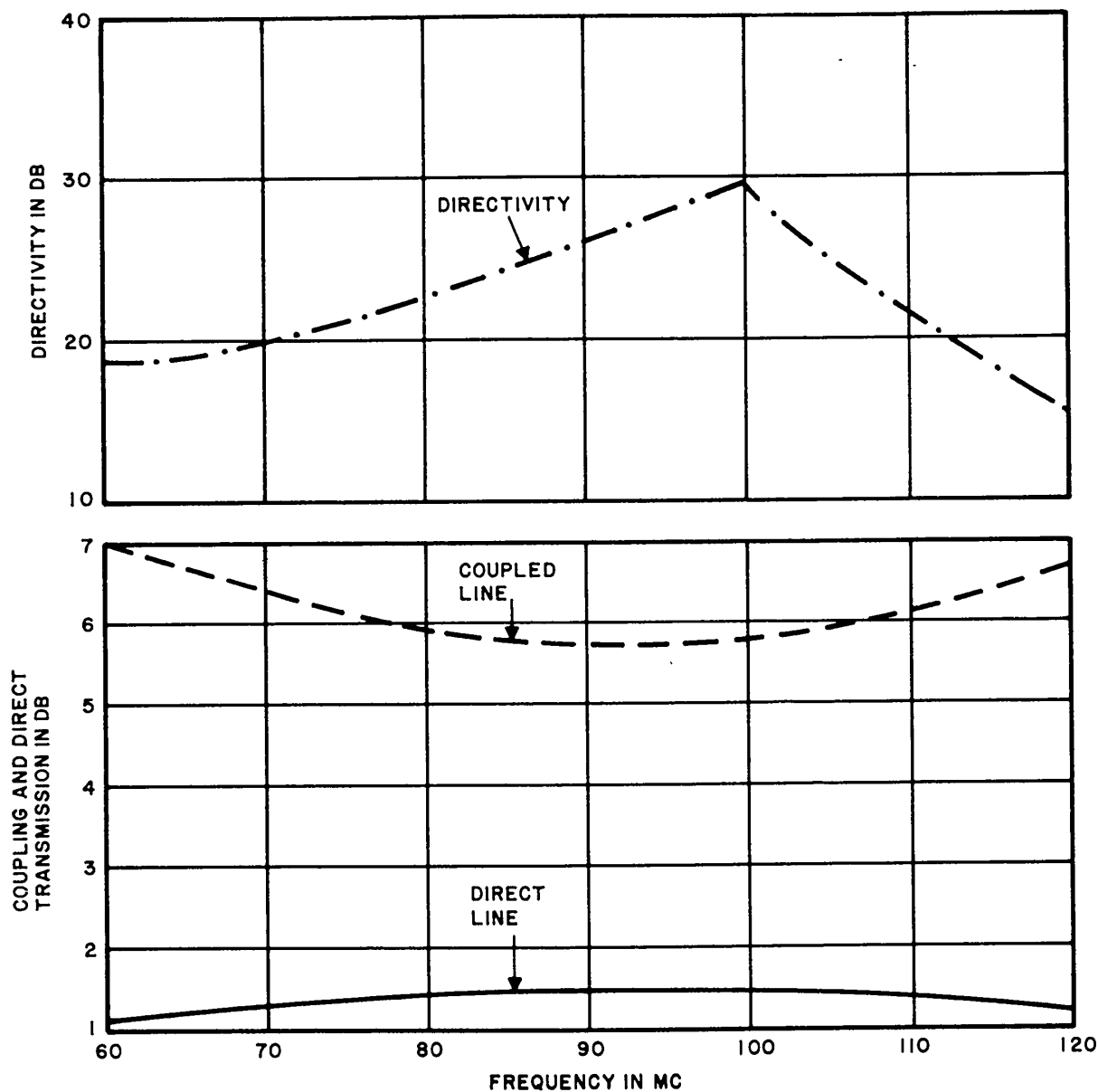


FIGURE 33. COUPLING, DIRECT TRANSMISSION, AND DIRECTIVITY VS FREQUENCY OF 6.0-DB DIRECTIONAL COUPLER FOR SUB-BAND 2

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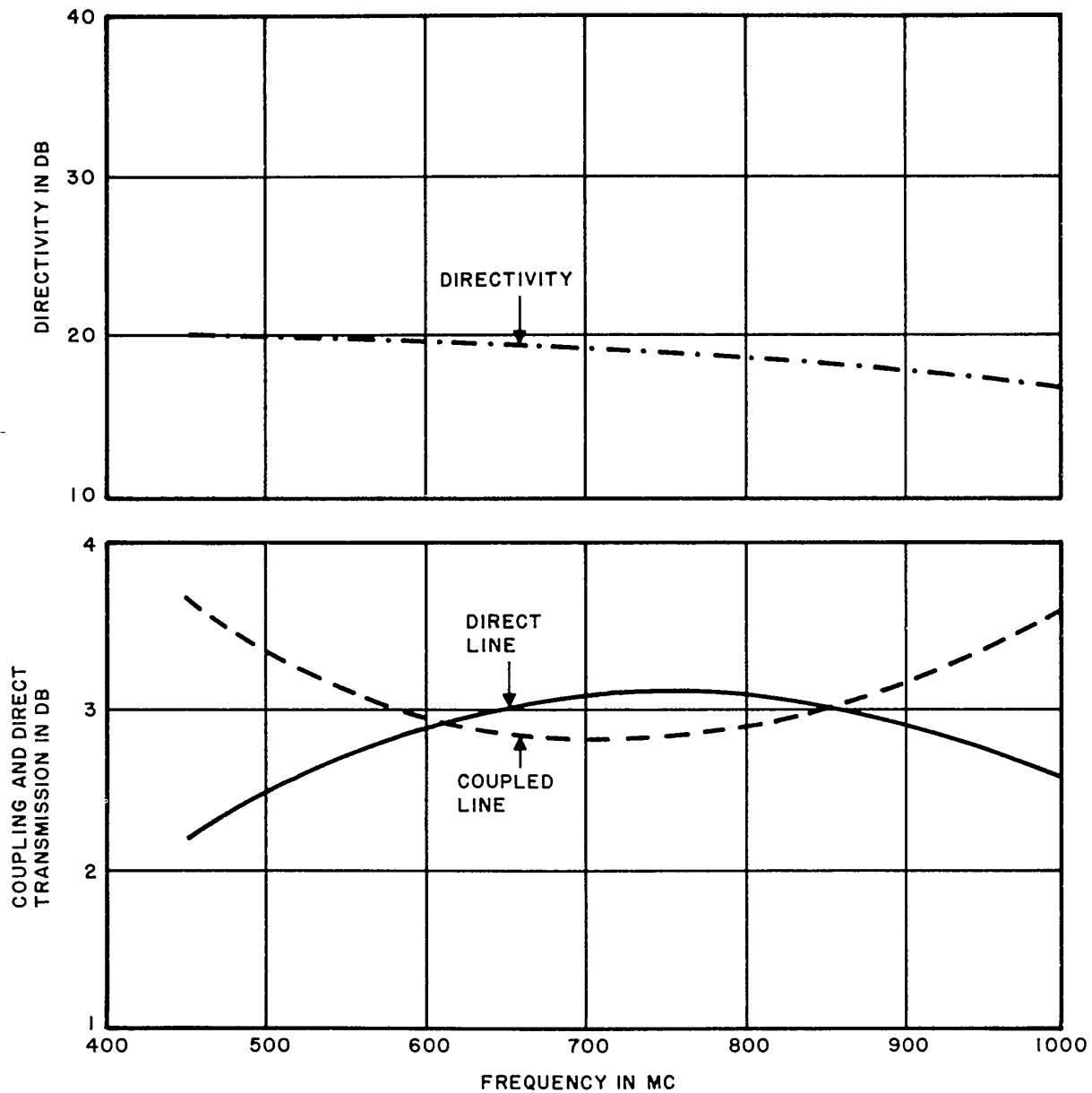


FIGURE 34. COUPLING, DIRECT TRANSMISSION, AND DIRECTIVITY VS FREQUENCY OF 3.0-DB DIRECTIONAL COUPLER FOR SUB-BAND 5

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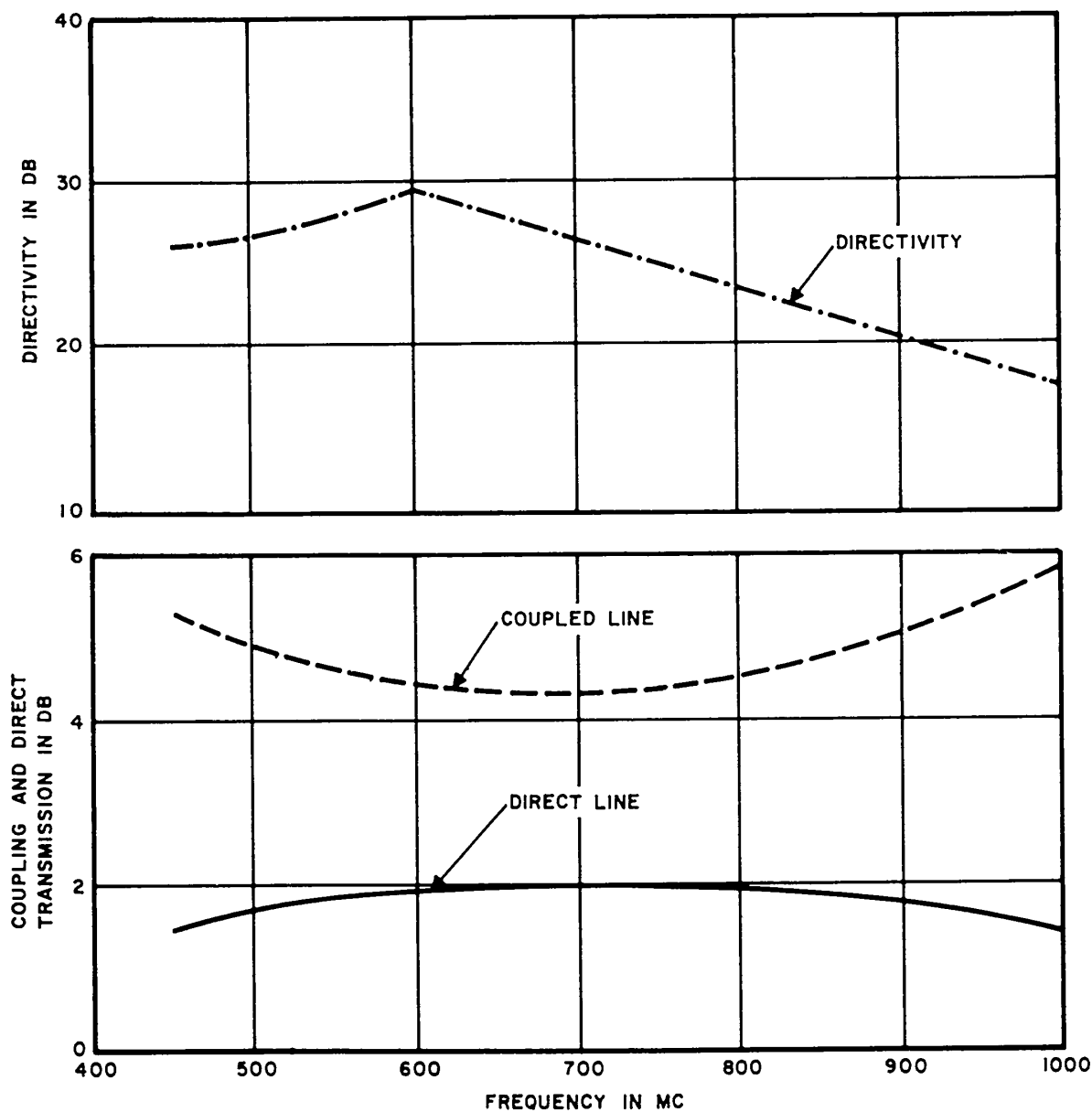


FIGURE 35. COUPLING, DIRECT TRANSMISSION, AND DIRECTIVITY VS FREQUENCY OF 4.8-DB DIRECTIONAL COUPLER FOR SUB-BAND 5

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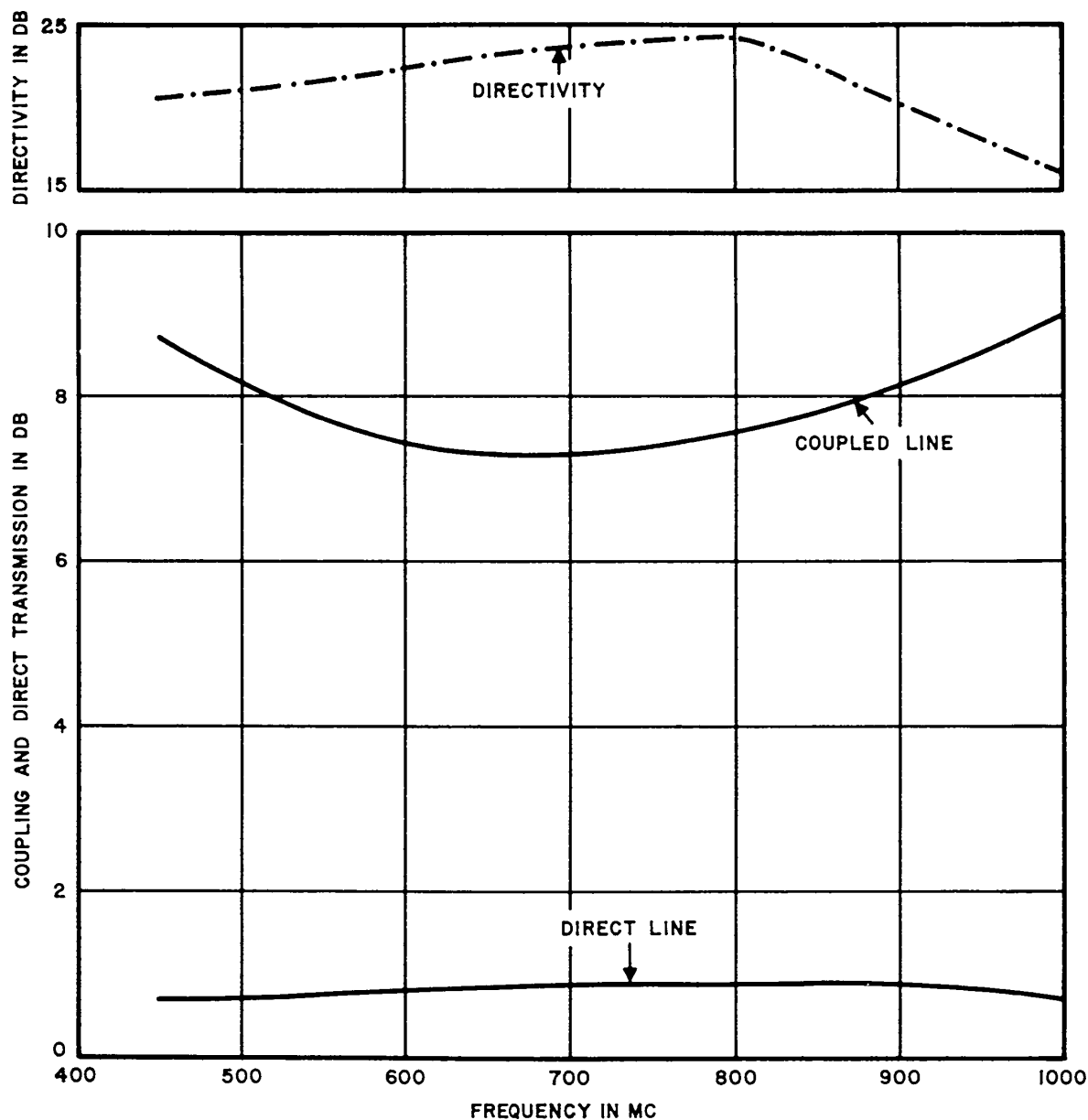


FIGURE 36. COUPLING, DIRECT TRANSMISSION, AND DIRECTIVITY VS FREQUENCY OF 7.5-DB DIRECTIONAL COUPLER FOR SUB-BAND 5

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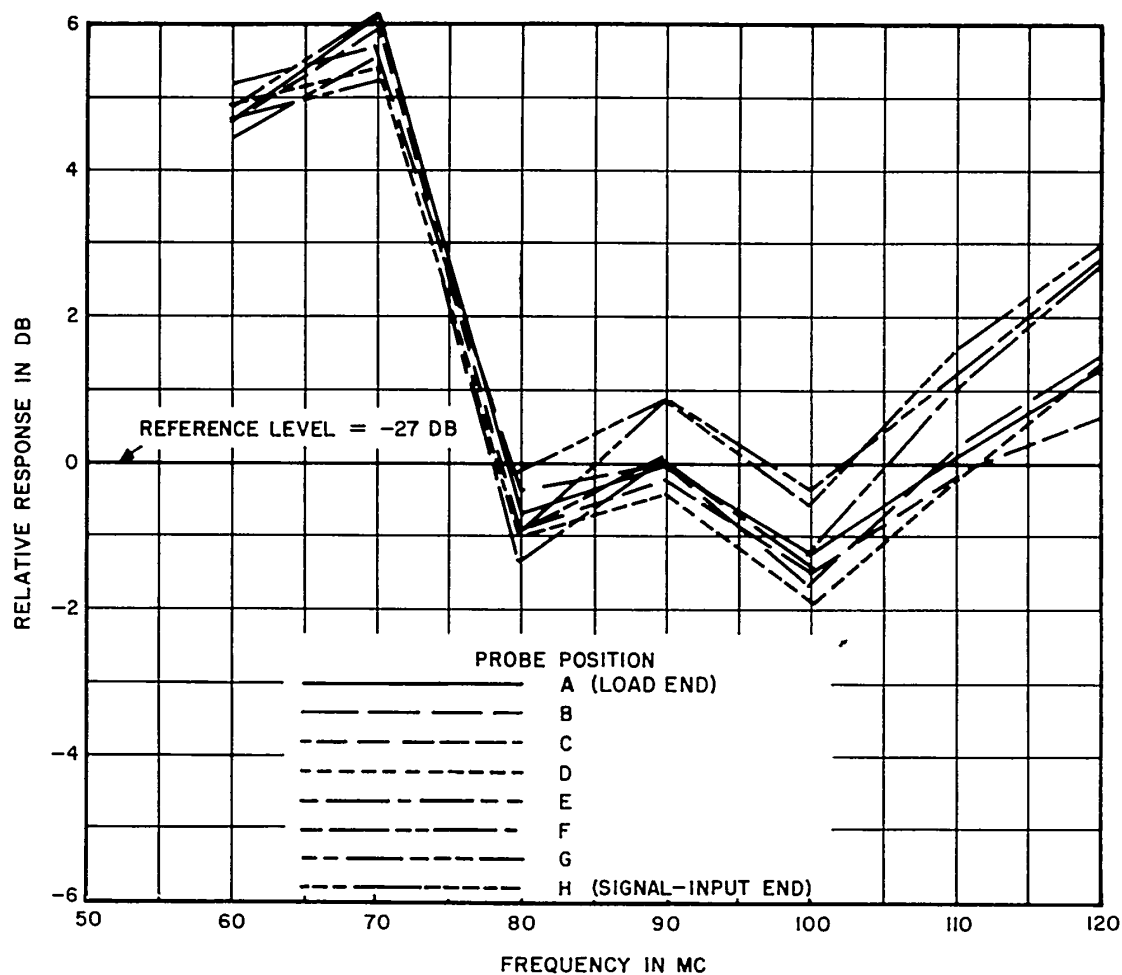


FIGURE 37. AMPLITUDE AND FREQUENCY CHARACTERISTICS OF PROBED LINE FOR SUB-BAND 2

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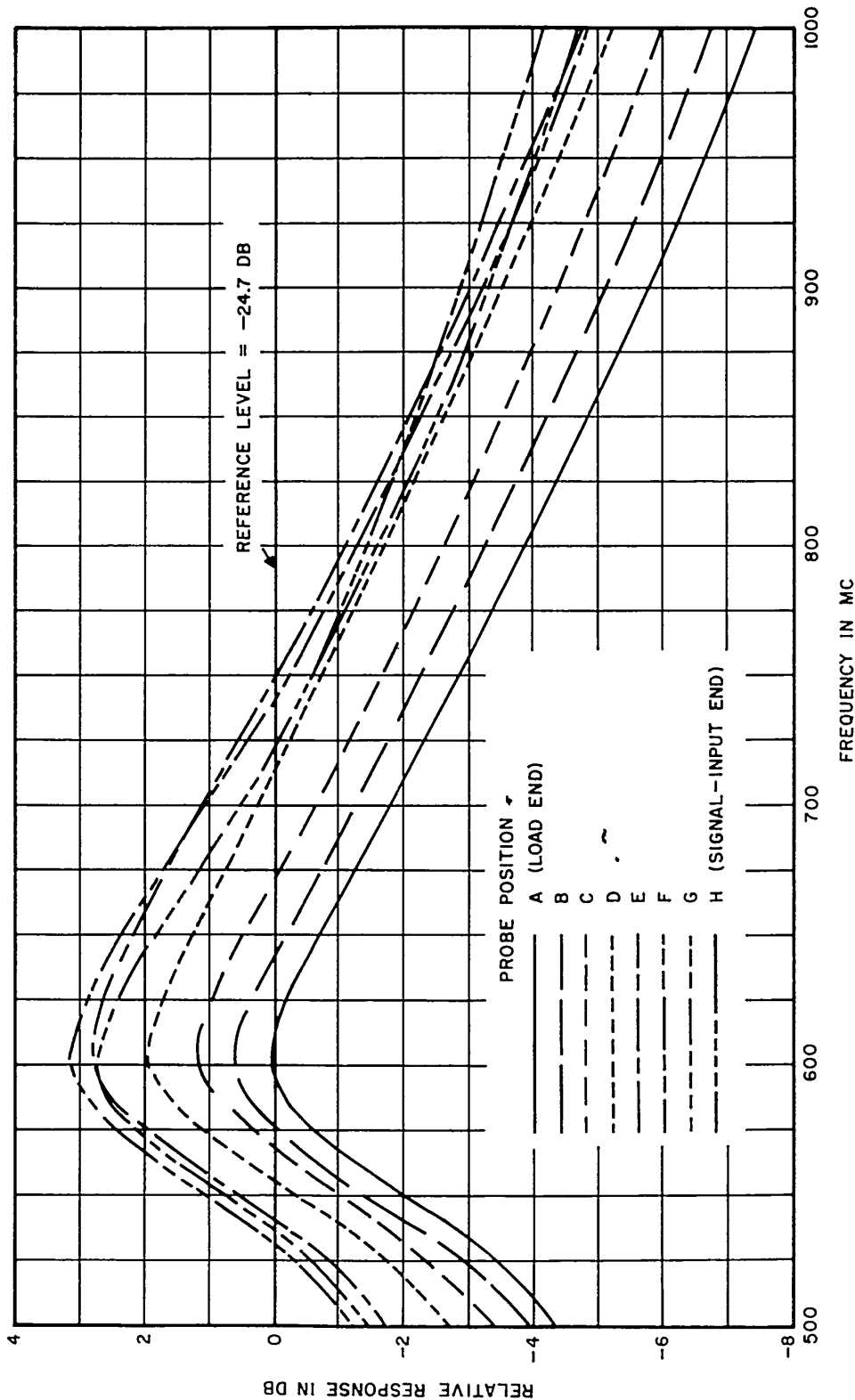


FIGURE 38. AMPLITUDE AND FREQUENCY CHARACTERISTICS OF PROBED LINE FOR SUB-BAND 5

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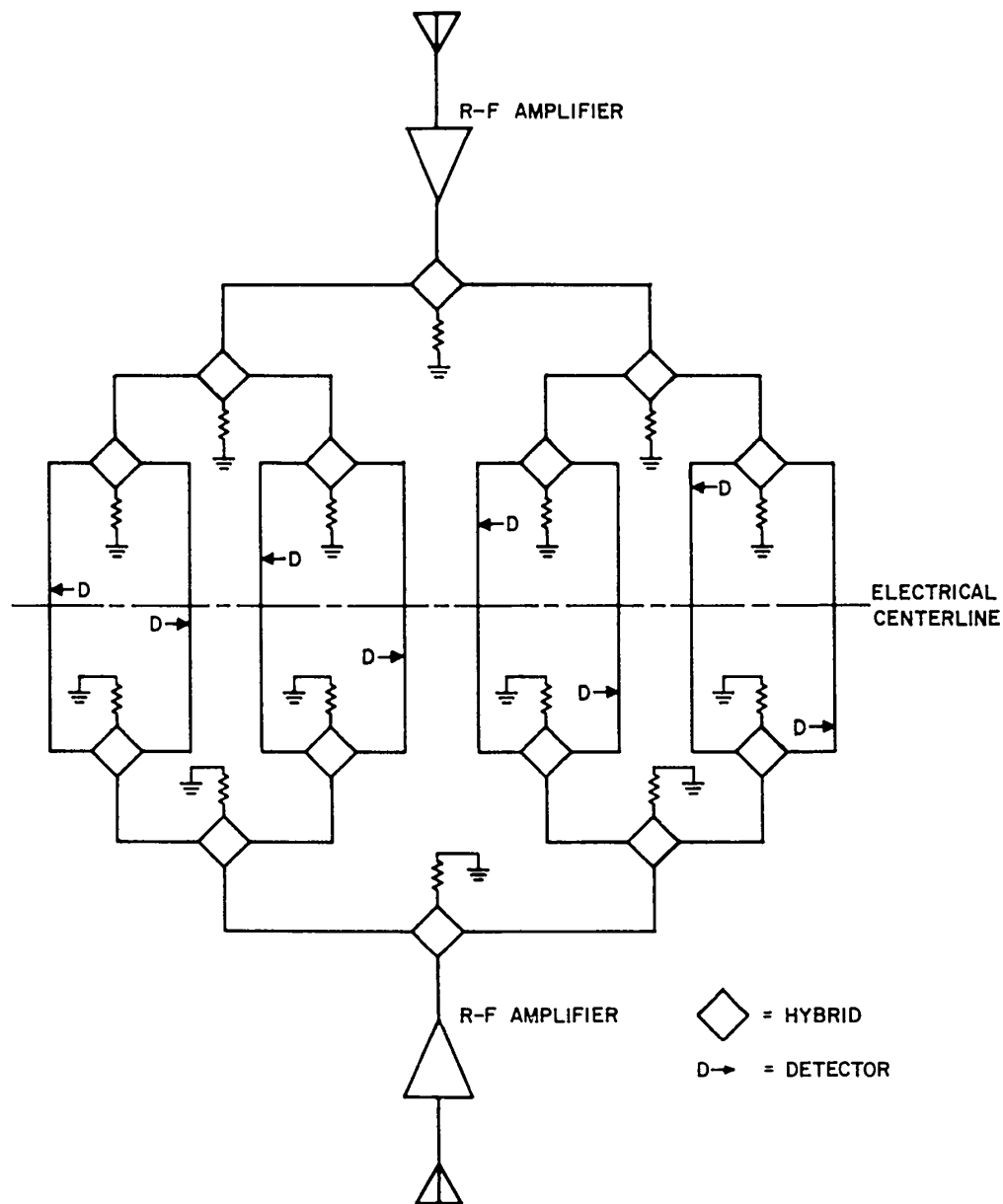


FIGURE 39. SIMPLIFIED BLOCK DIAGRAM OF DIRECTION ANALYZER USING HYBRID-BINARY-TREE ARRANGEMENT OF PROBED LINES

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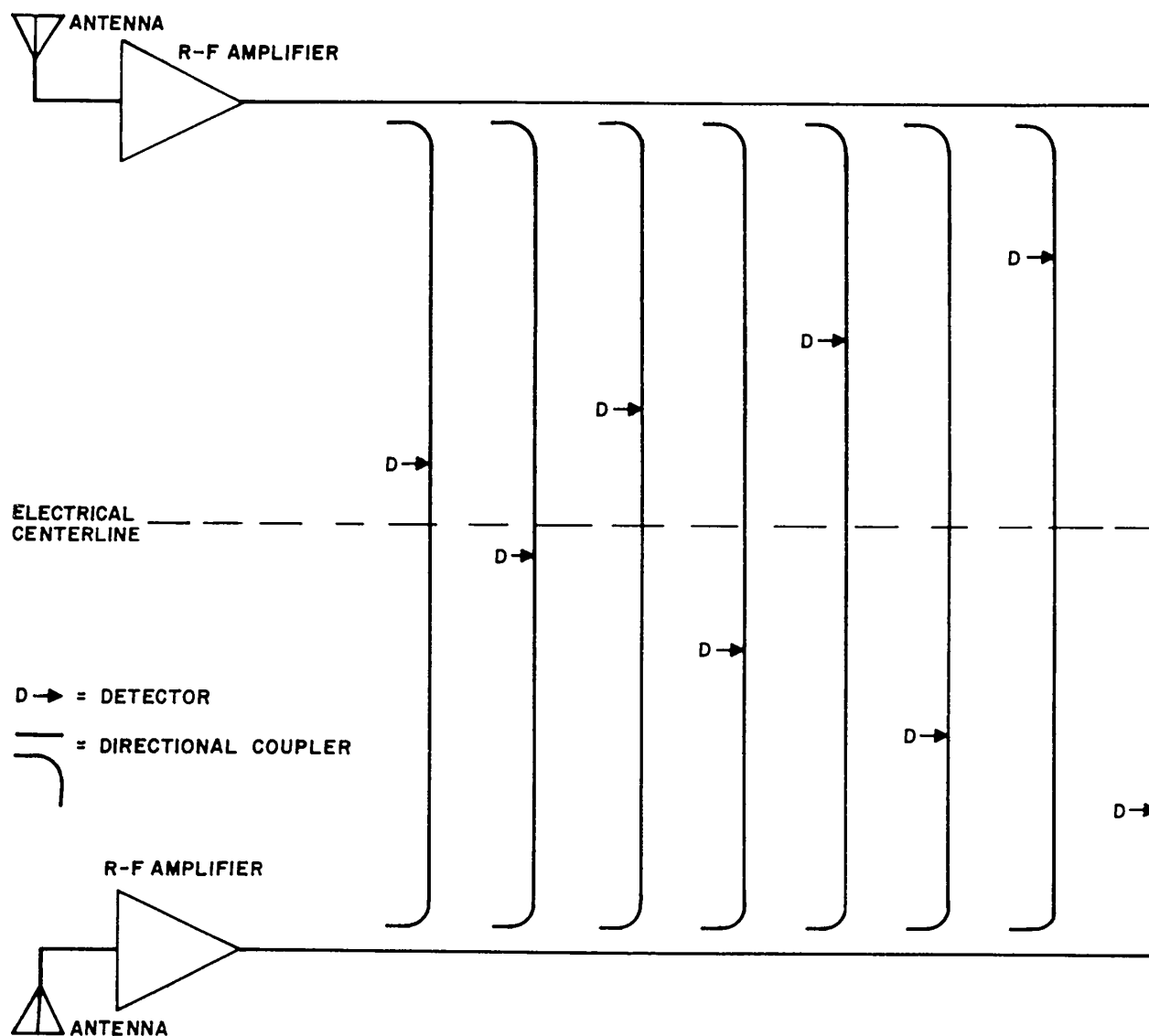


FIGURE 40. SIMPLIFIED BLOCK DIAGRAM OF DIRECTION ANALYZER USING DIRECTIONAL-COUPLER FEED TO PROBED LINES

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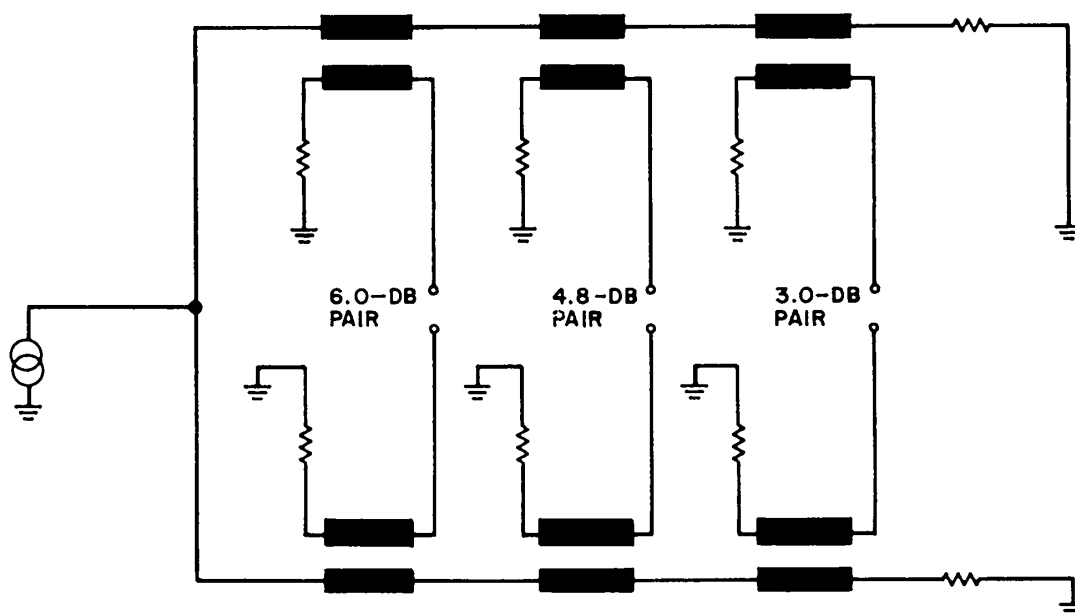


FIGURE 4I. SIMPLIFIED SCHEMATIC DIAGRAM OF EXPERIMENTAL ASSEMBLY OF DIRECTIONAL COUPLERS FOR SUB-BAND 2

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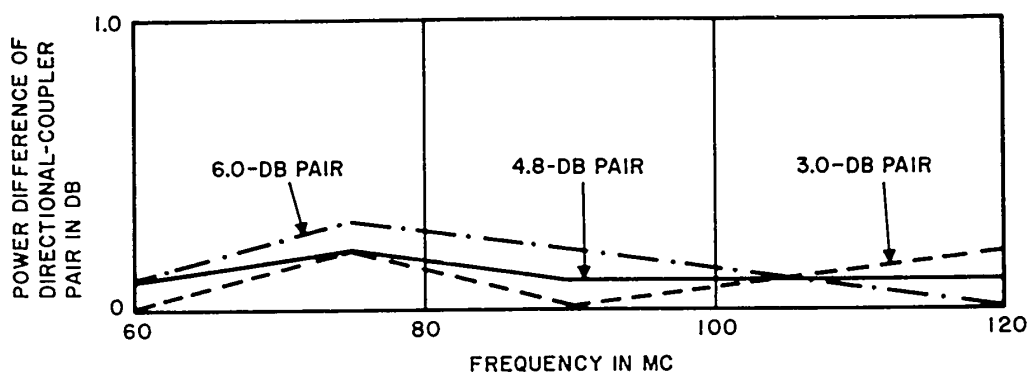


FIGURE 42. POWER MATCH OF DIRECTIONAL COUPLERS

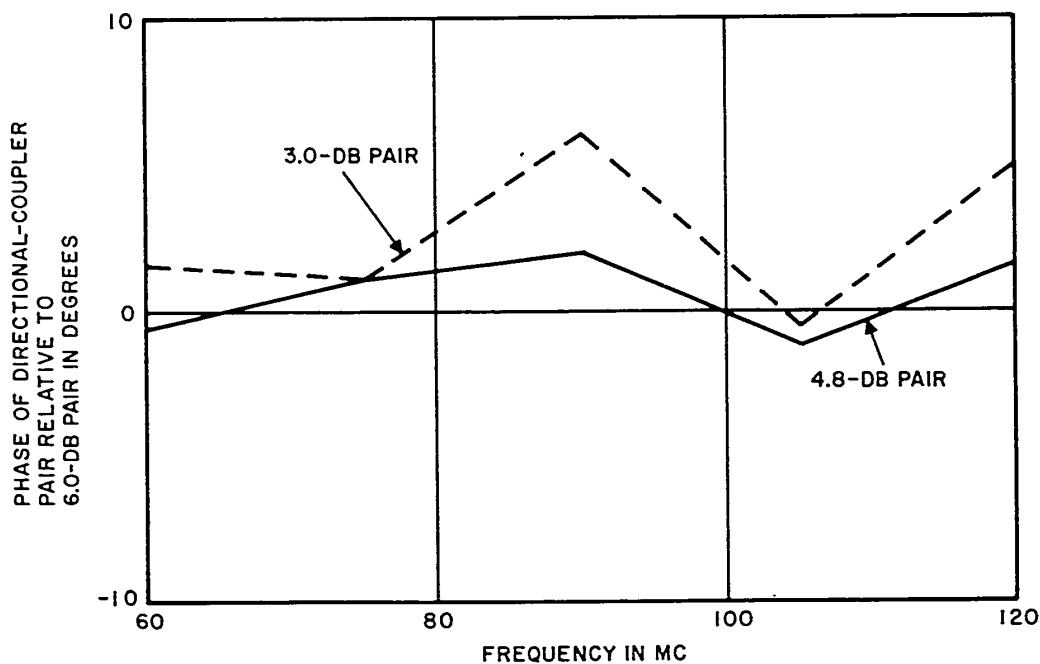


FIGURE 43. PHASE TRACKING OF DIRECTIONAL-COUPLER PAIRS

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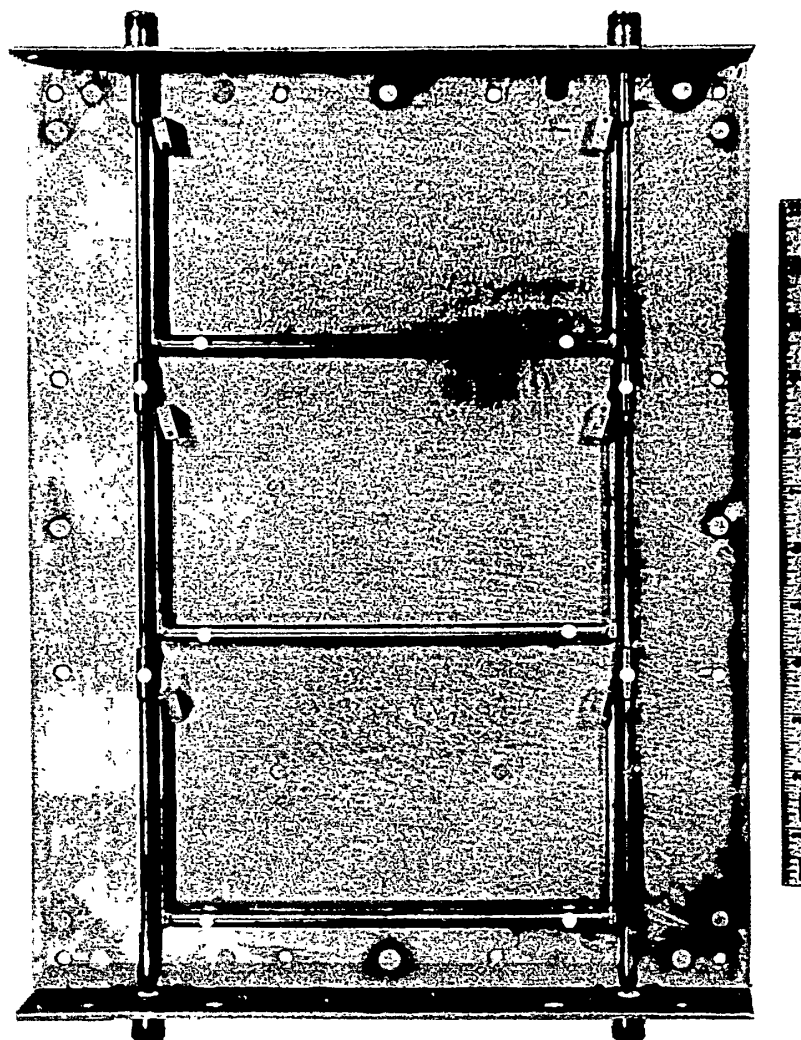


FIGURE 44. INTERNAL VIEW OF EXPERIMENTAL ASSEMBLY OF DIRECTIONAL COUPLERS FOR SUB-BAND 5

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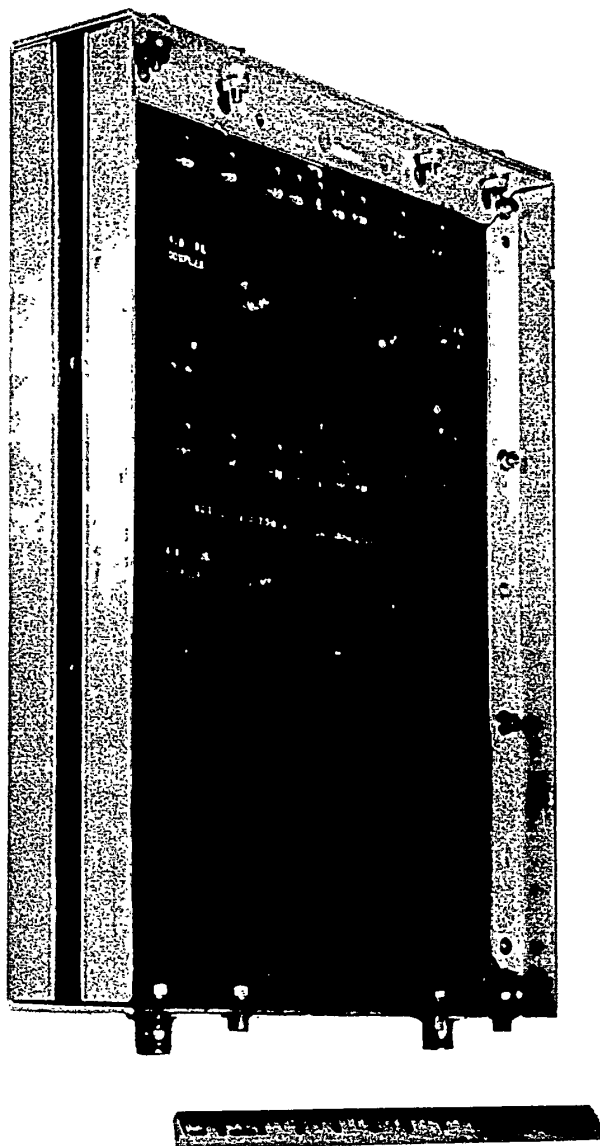


FIGURE 45. ASSEMBLY OF FIGURE 44 WITH COVER IN PLACE

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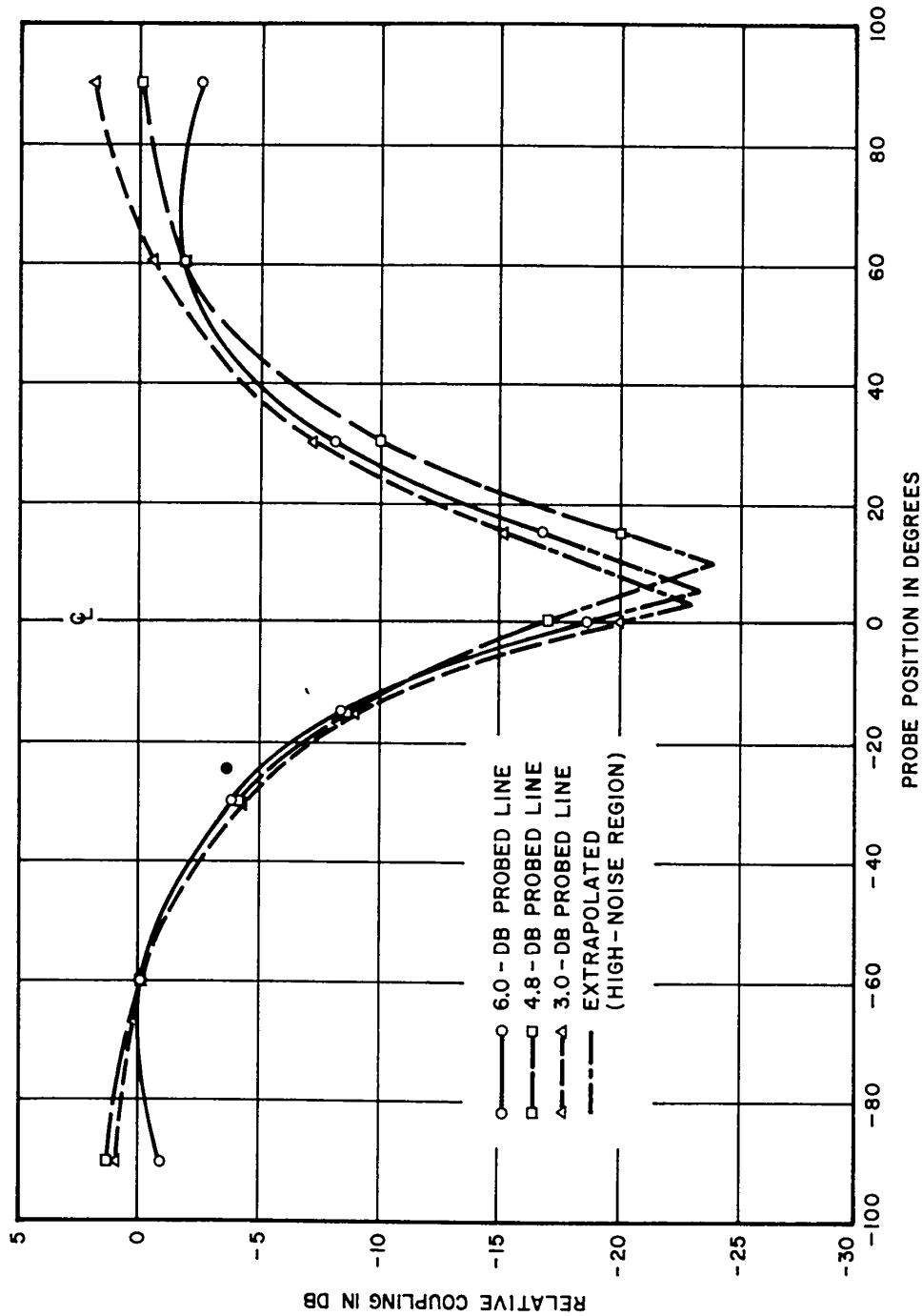


FIGURE 46. SWR PATTERNS OF ASSEMBLY OF FIGURE 44 AT 500 MC

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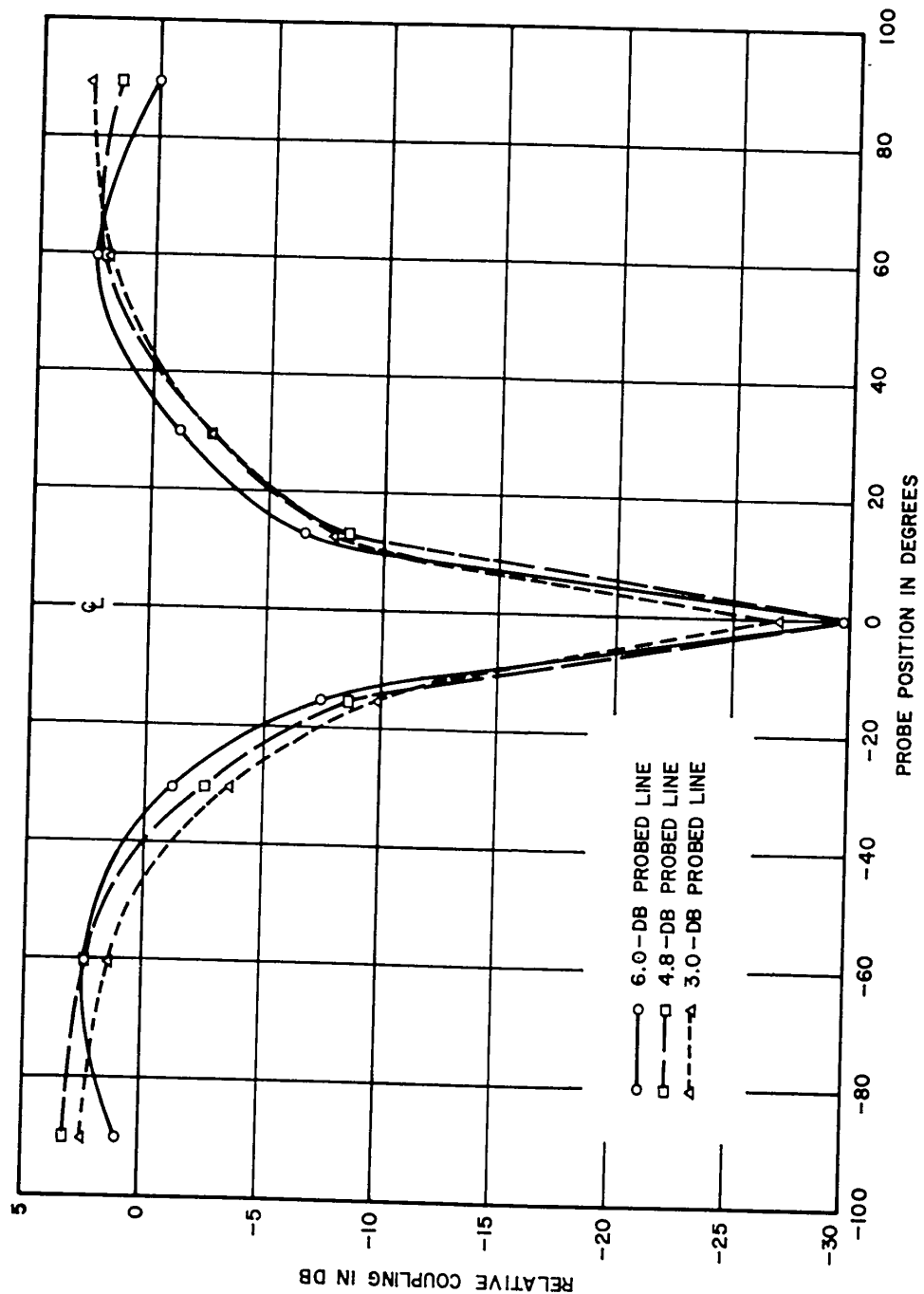


FIGURE 47. SWR PATTERNS OF ASSEMBLY OF FIGURE 44 AT 750 MC

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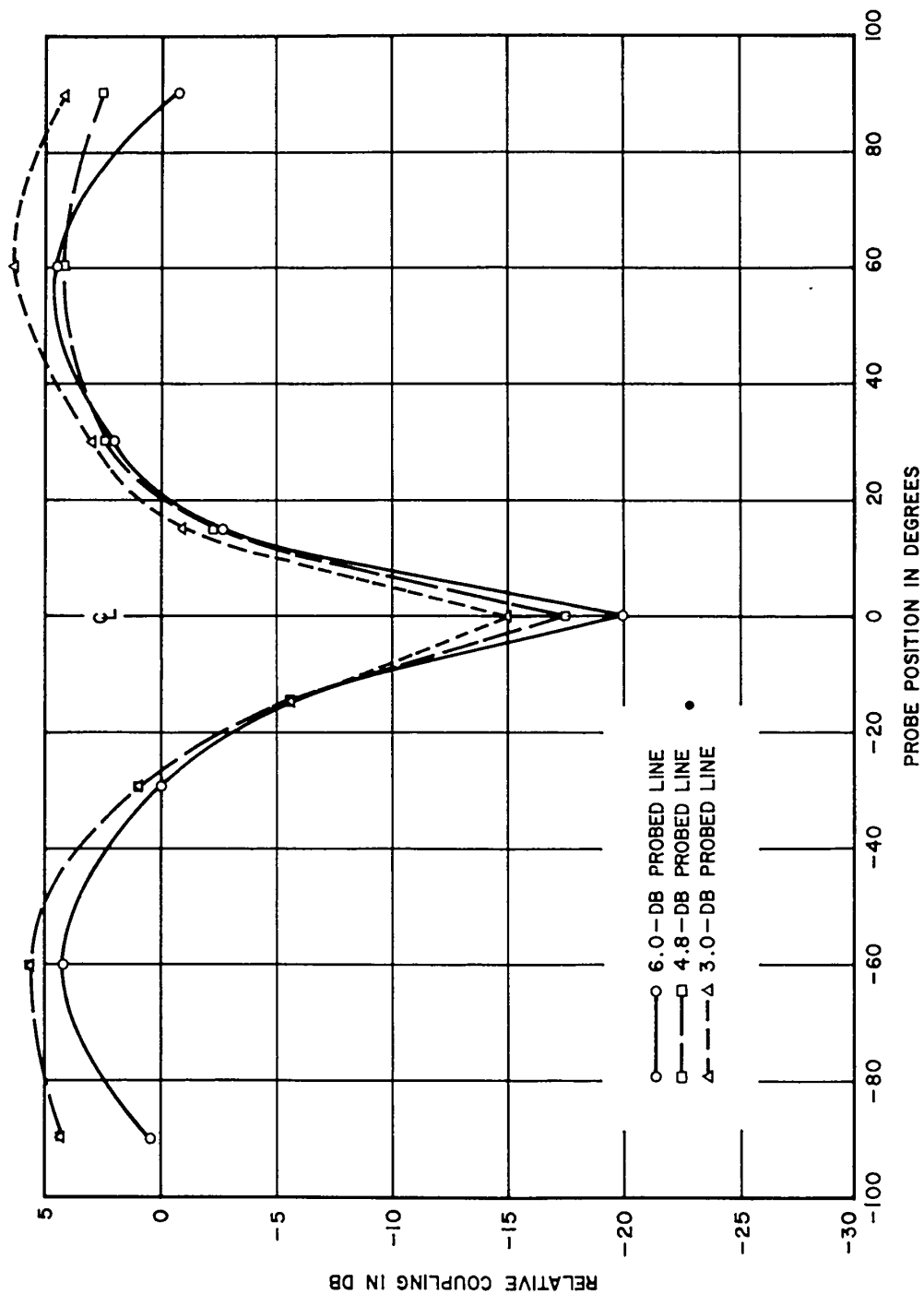


FIGURE 48. SWR PATTERNS OF ASSEMBLY OF FIGURE 44 AT 1000 MC

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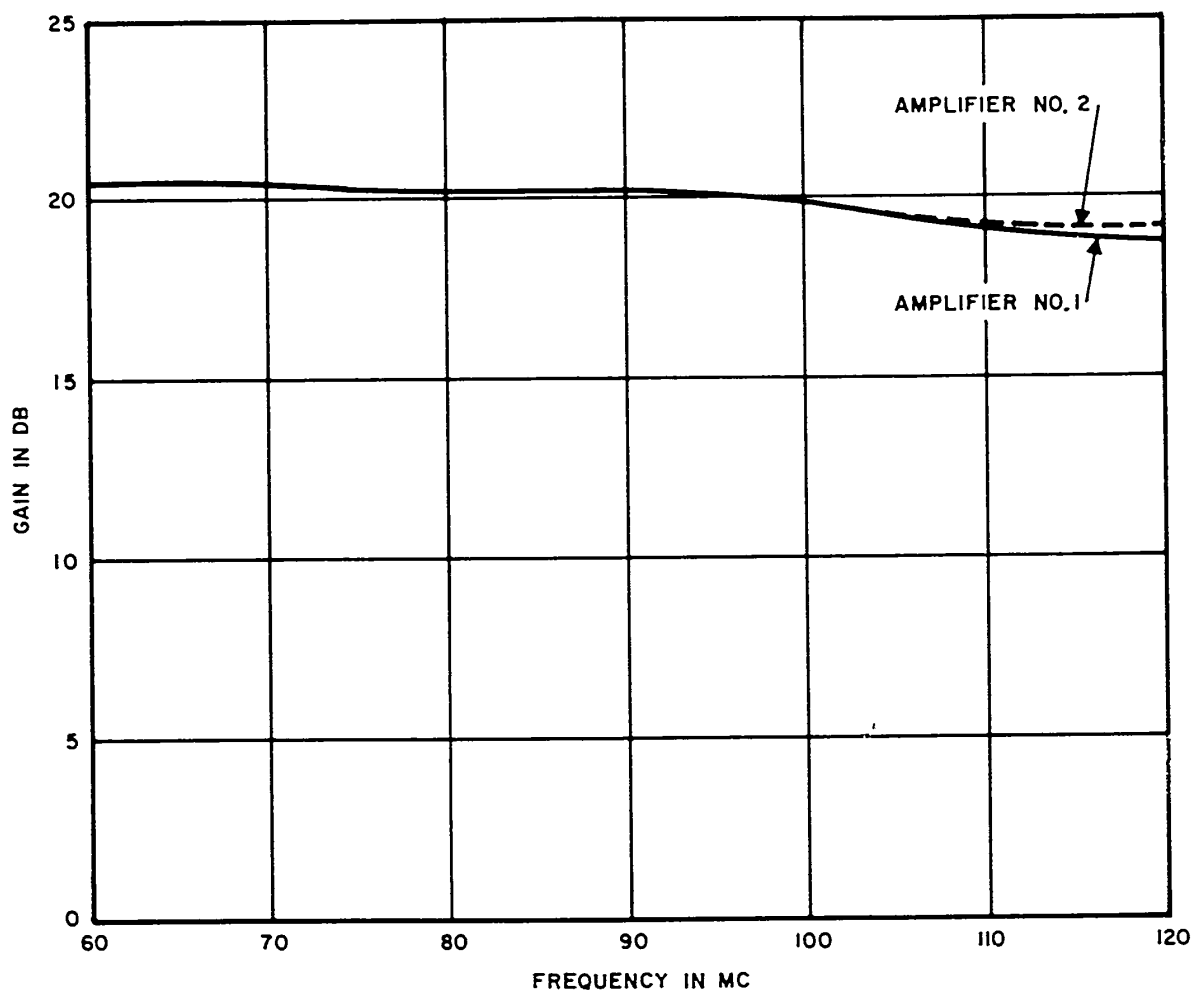


FIGURE 49. GAIN VS FREQUENCY OF DISTRIBUTED AMPLIFIERS FOR SUB-BAND 2

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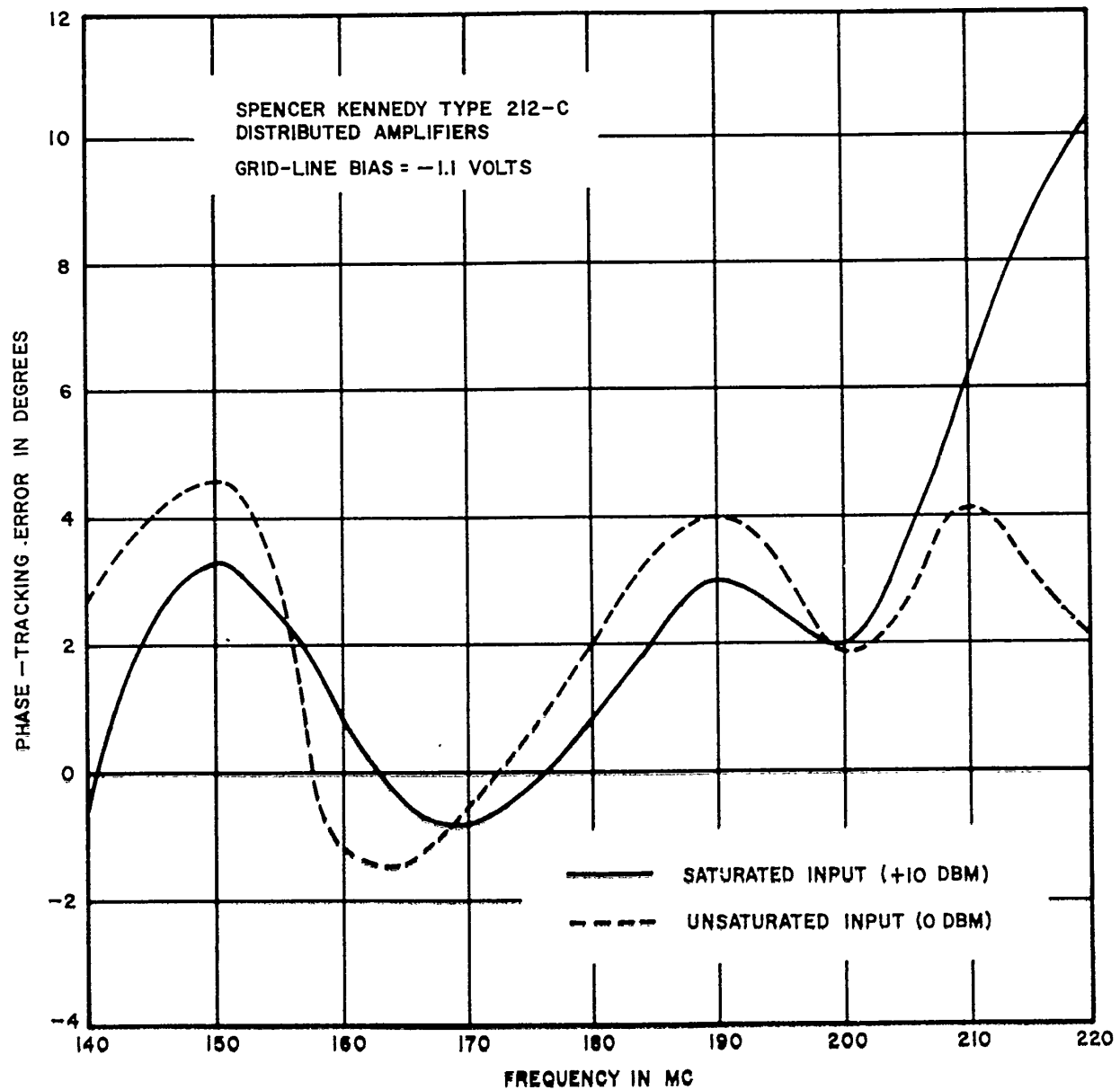


FIGURE 50. PHASE-TRACKING ERROR IN DISTRIBUTED AMPLIFIERS

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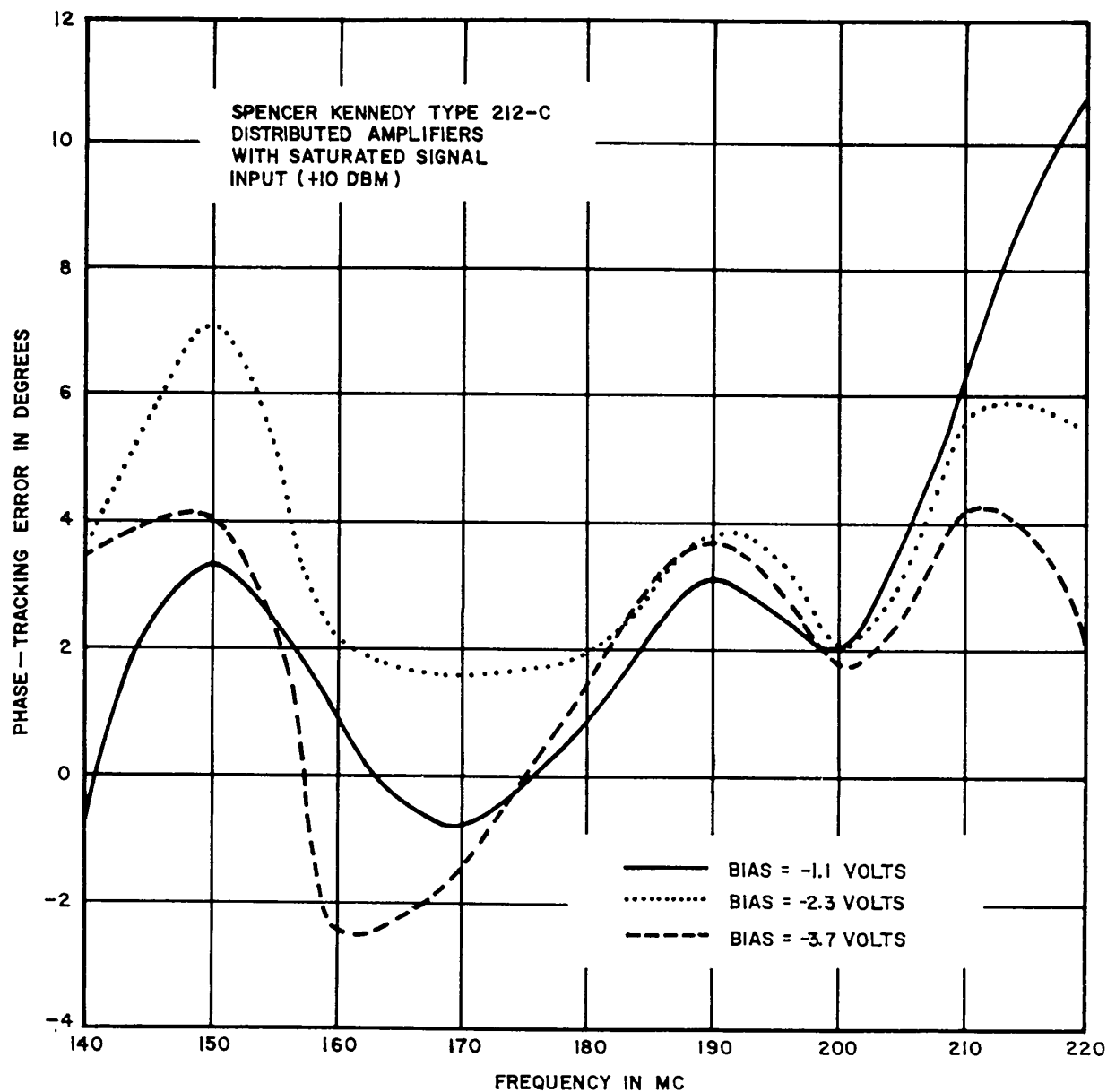


FIGURE 5I. EFFECT OF GAIN CONTROL ON PHASE-TRACKING
ERROR IN SATURATION REGION

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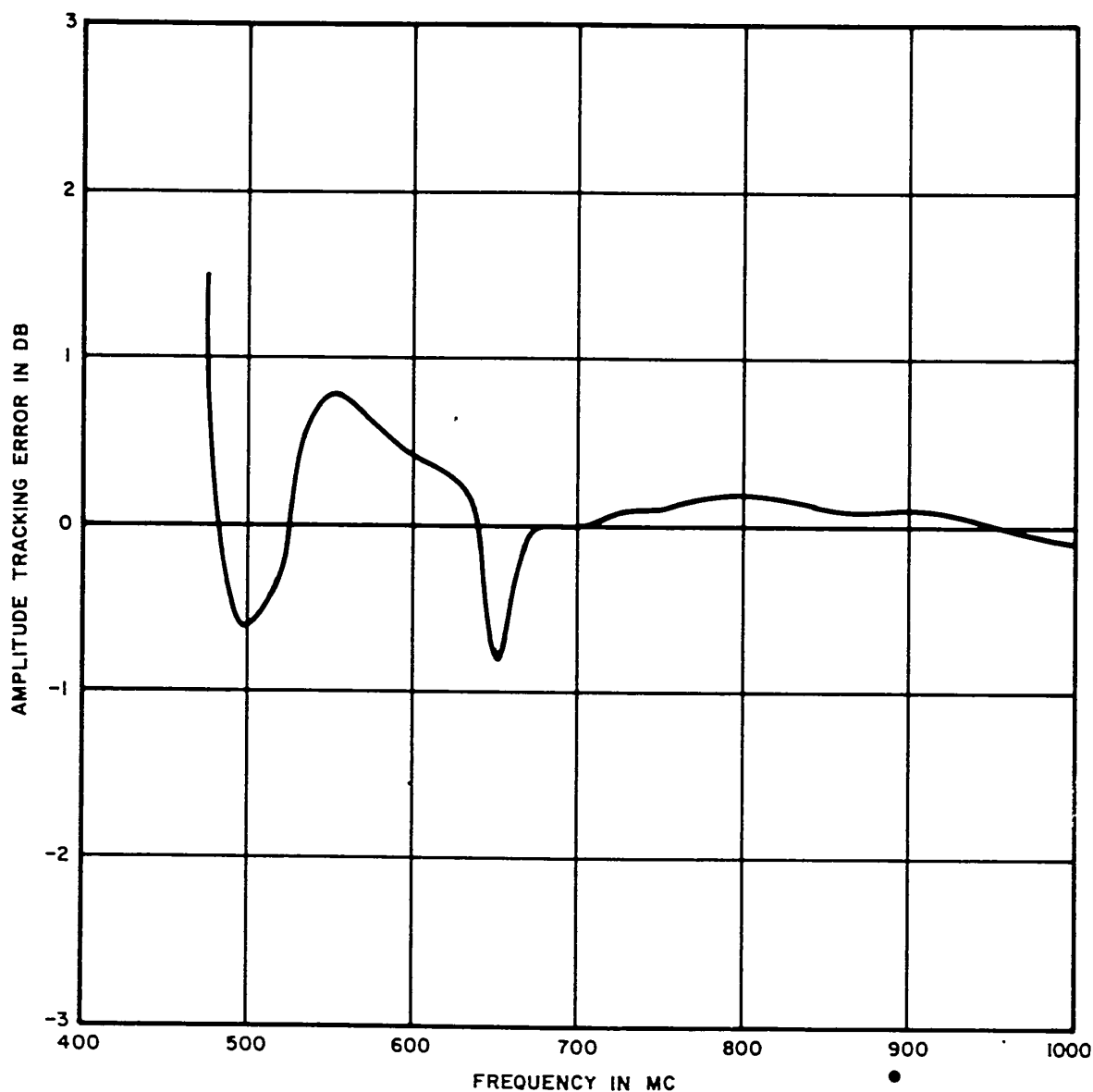


FIGURE 52. AMPLITUDE-TRACKING ERROR VS FREQUENCY OF TWT'S
FOR SUB-BAND 5

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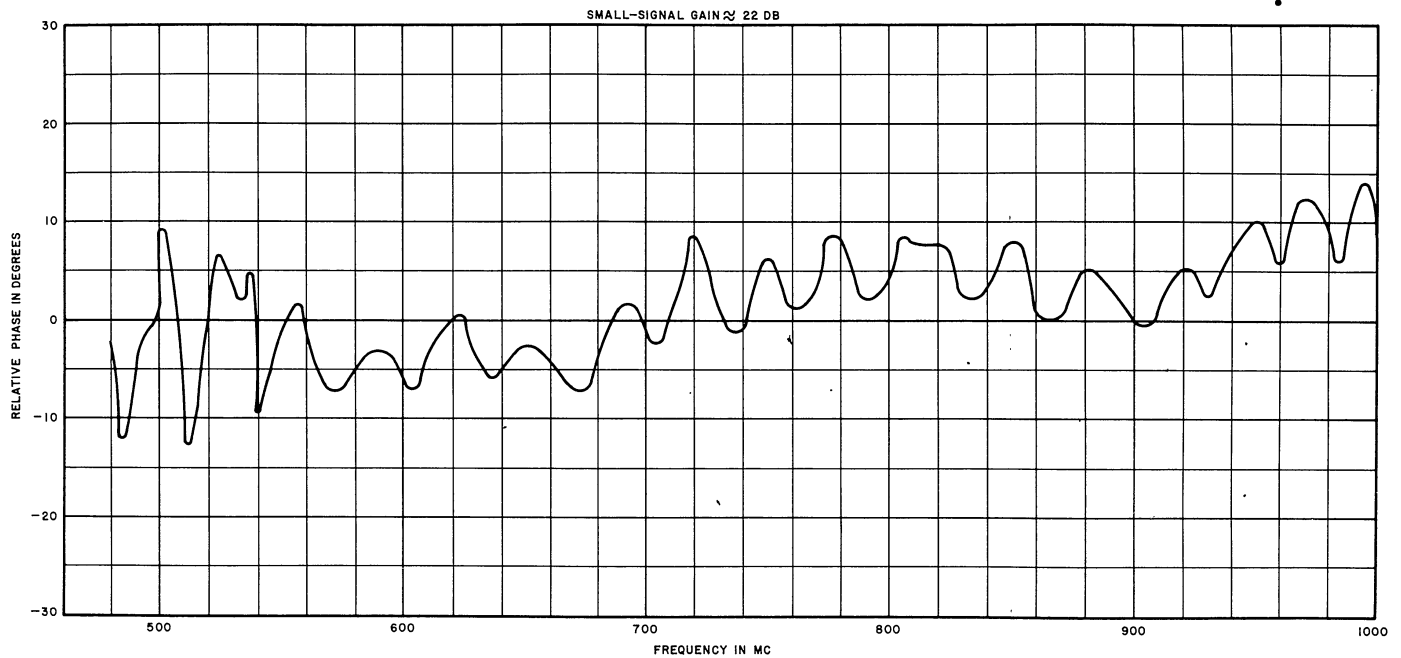


FIGURE 53. PHASE-TRACKING ERROR VS FREQUENCY OF TWT'S FOR SUB-BAND 5

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FIGURE 53
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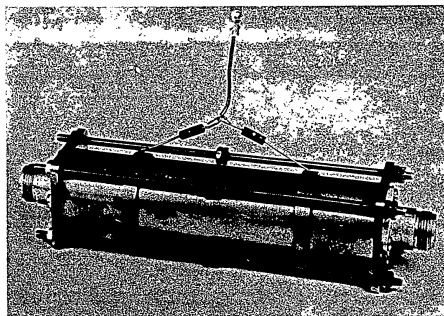
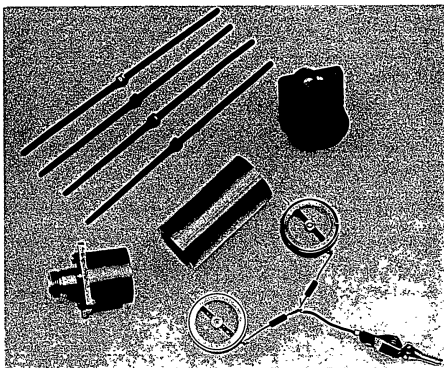


FIGURE 54. EXPERIMENTAL FAST-ACTING ATTENUATOR, DISASSEMBLED

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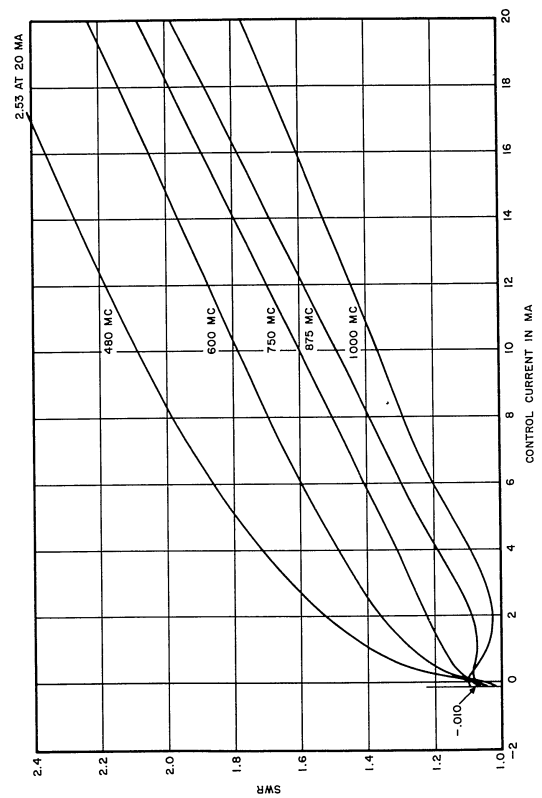


FIGURE 55. SWR CHARACTERISTICS OF EXPERIMENTAL FAST-ACTING ATTENUATOR

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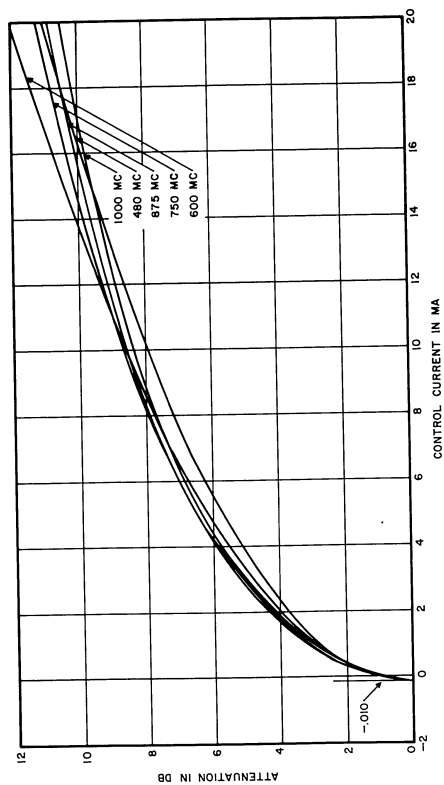


FIGURE 56. ATTENUATION CHARACTERISTICS OF EXPERIMENTAL FAST-ACTING ATTENUATOR

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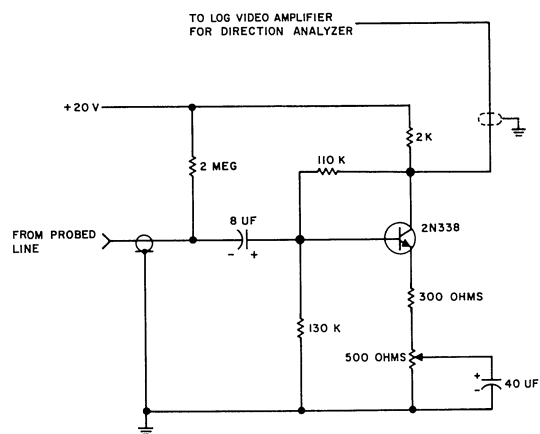


FIGURE 57. SCHEMATIC DIAGRAM OF PREAMPLIFIER FOR DIRECTION ANALYZER

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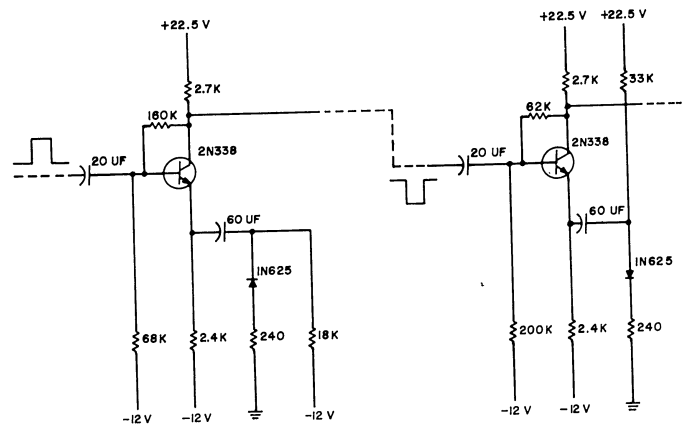


FIGURE 58. SCHEMATIC DIAGRAM OF TYPICAL GAIN-SWITCHED STAGES

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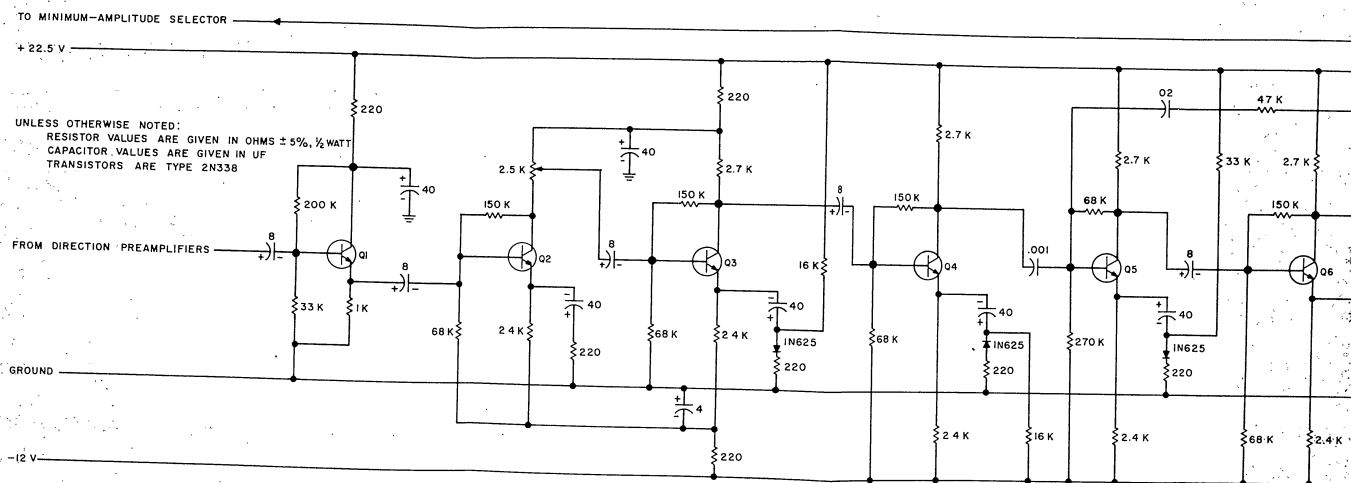


FIGURE 59. SC

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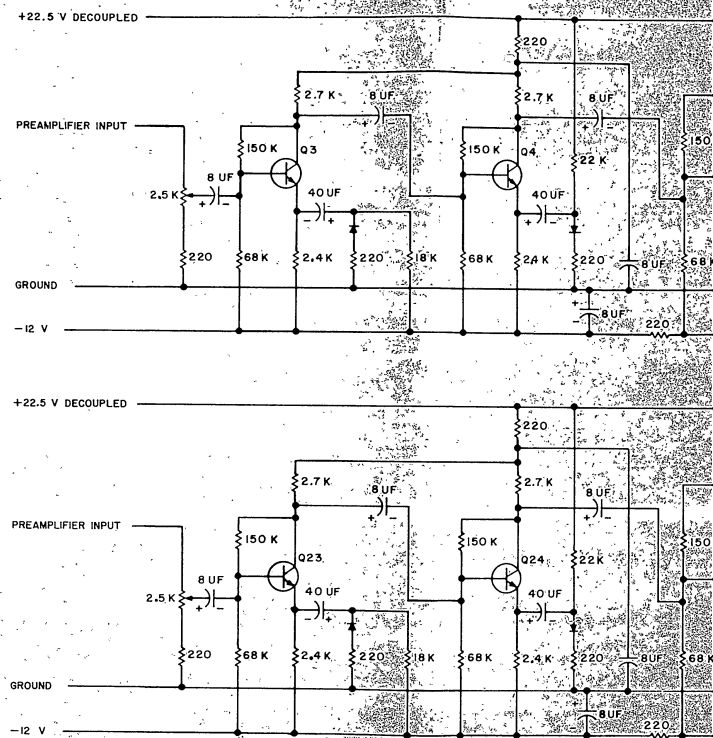
- 249



FIGURE 6L. ACTIVE AND PASSIVE DIFFERENCE CIRCUITS OF ONE CHANNEL OF MINIMUM-AMPLITUDE SELECTOR

FIGURE 62. SIMPLIFIED SCHEMATIC DIAGRAM OF MINIMUM-AMPLITUDE SELECTOR

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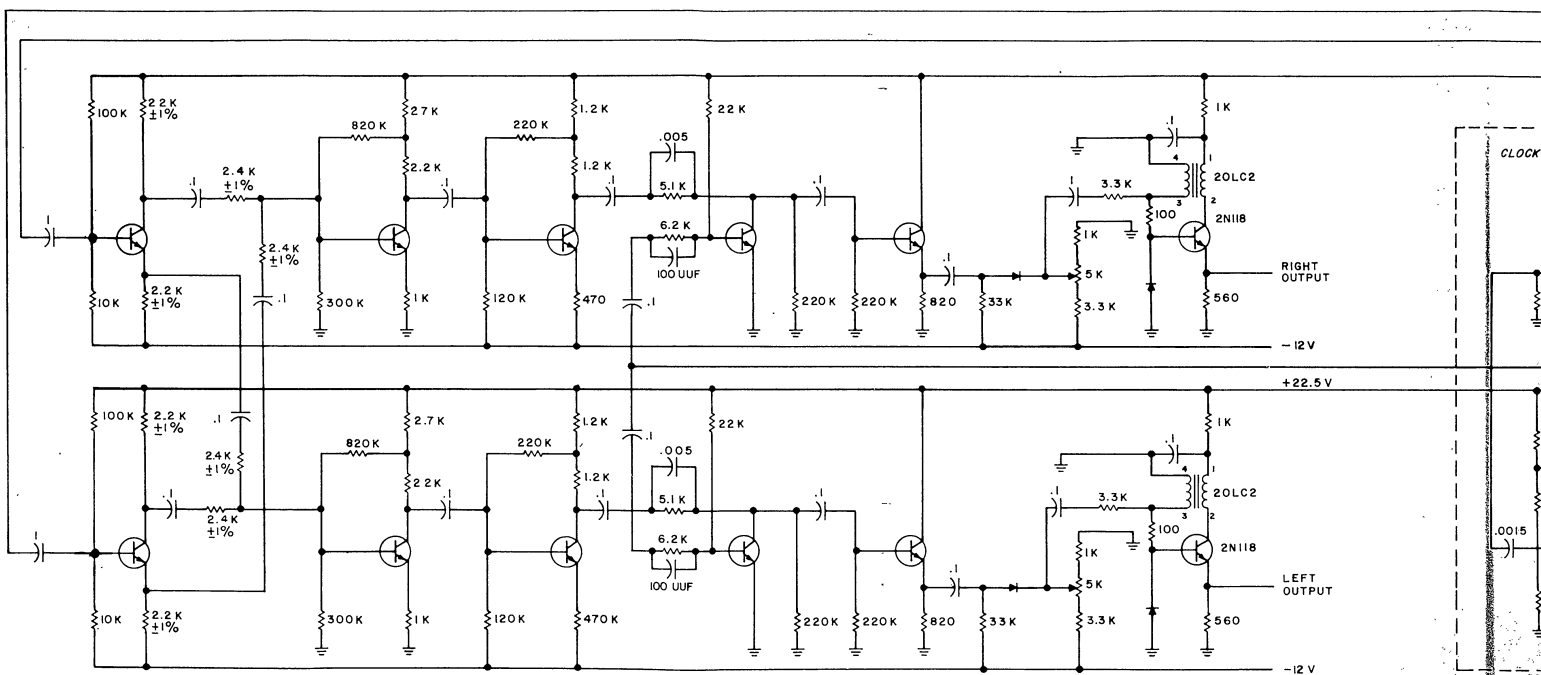
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FIGURE 64
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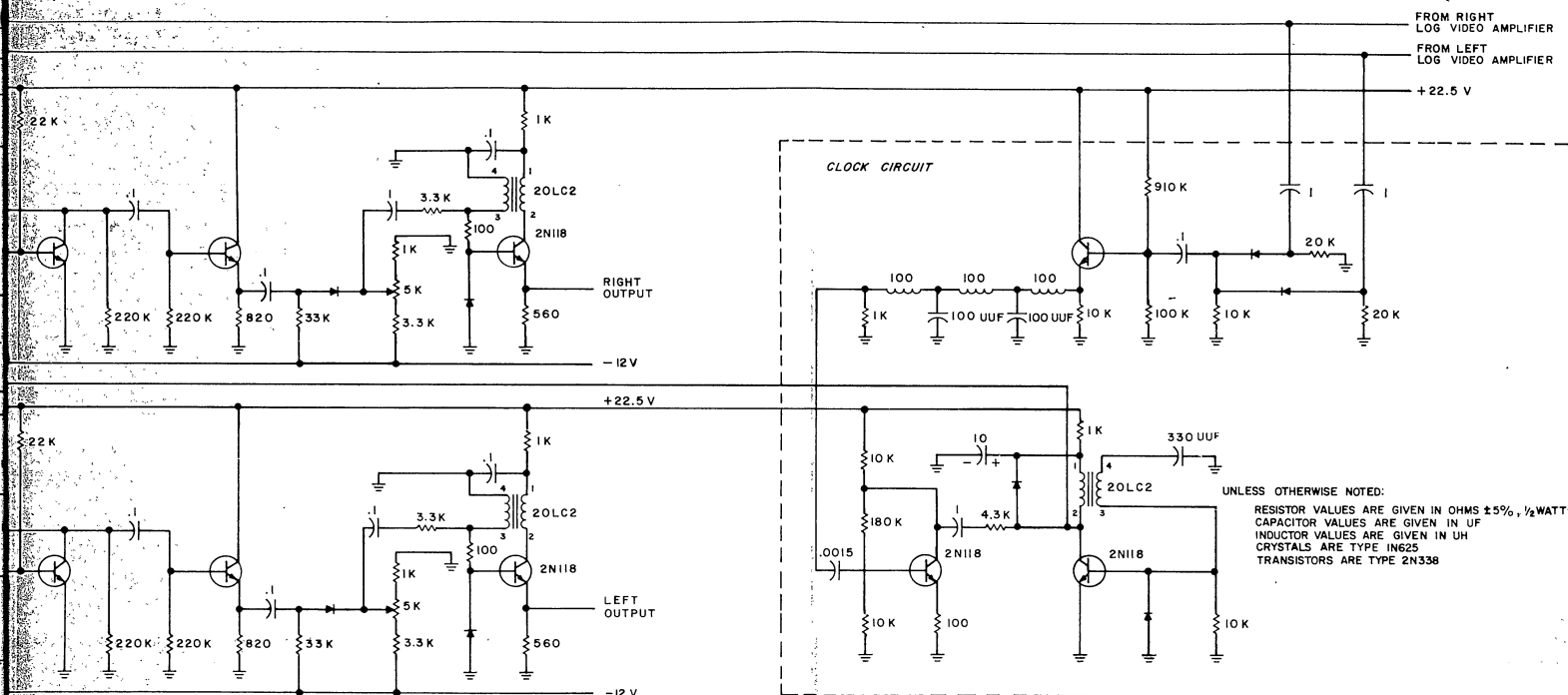


FIGURE 65. SCHEMATIC DIAGRAM OF PEAK-AMPLITUDE SELECTOR FOR LEFT-RIGHT INDICATOR

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FIGURE 65

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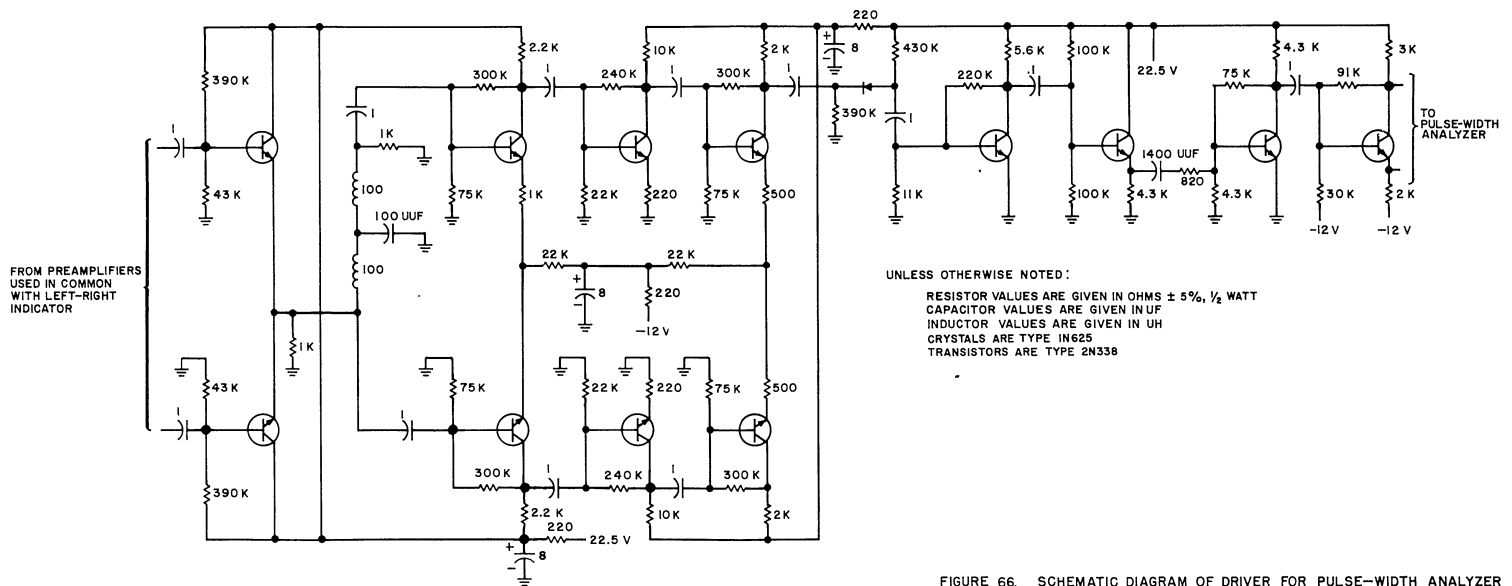


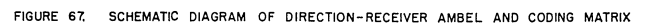
FIGURE 66. SCHEMATIC DIAGRAM OF DRIVER FOR PULSE-WIDTH ANALYZER

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FIGURE 66
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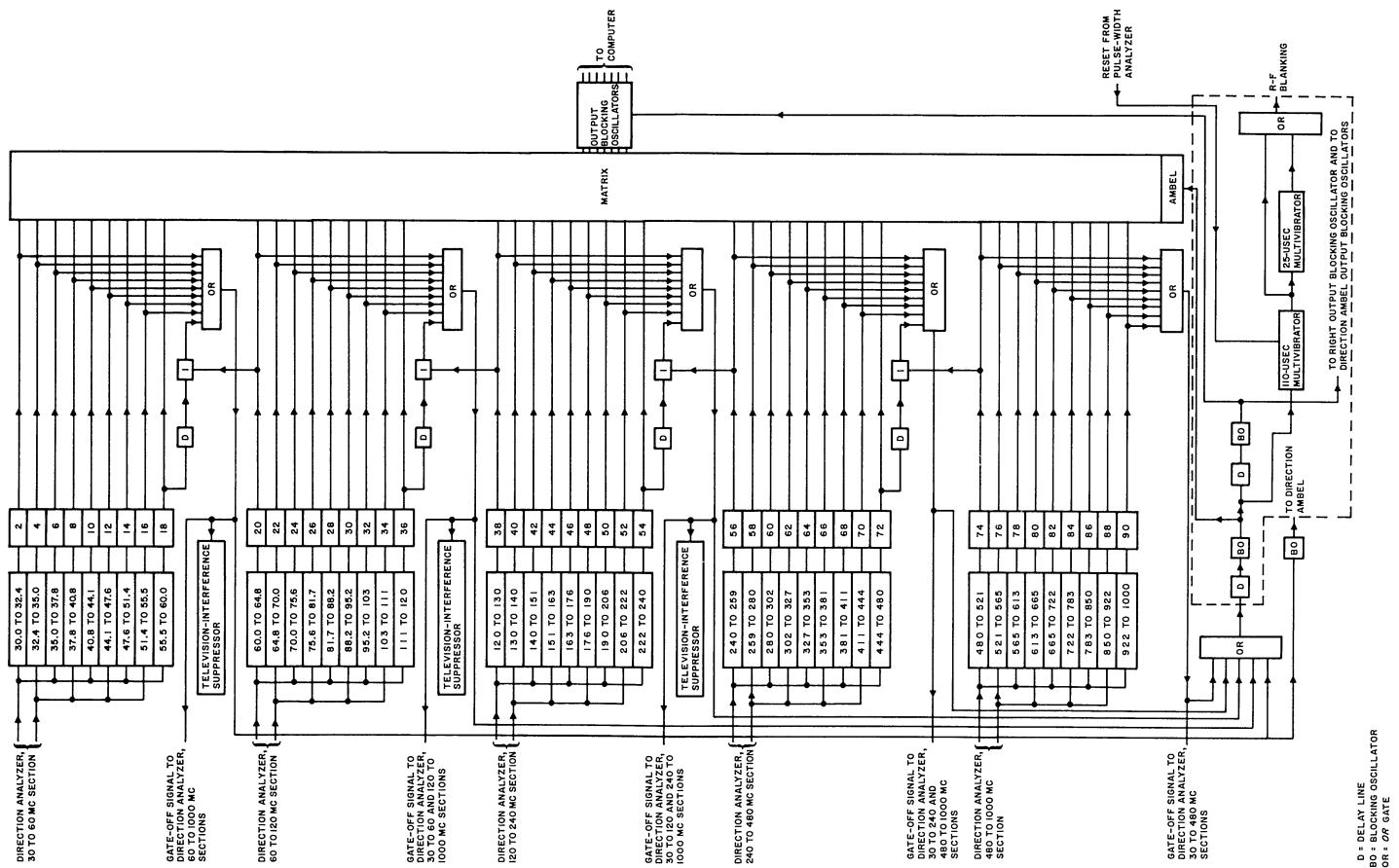


FIGURE 68. BLOCK DIAGRAM OF FREQUENCY ANALYZER

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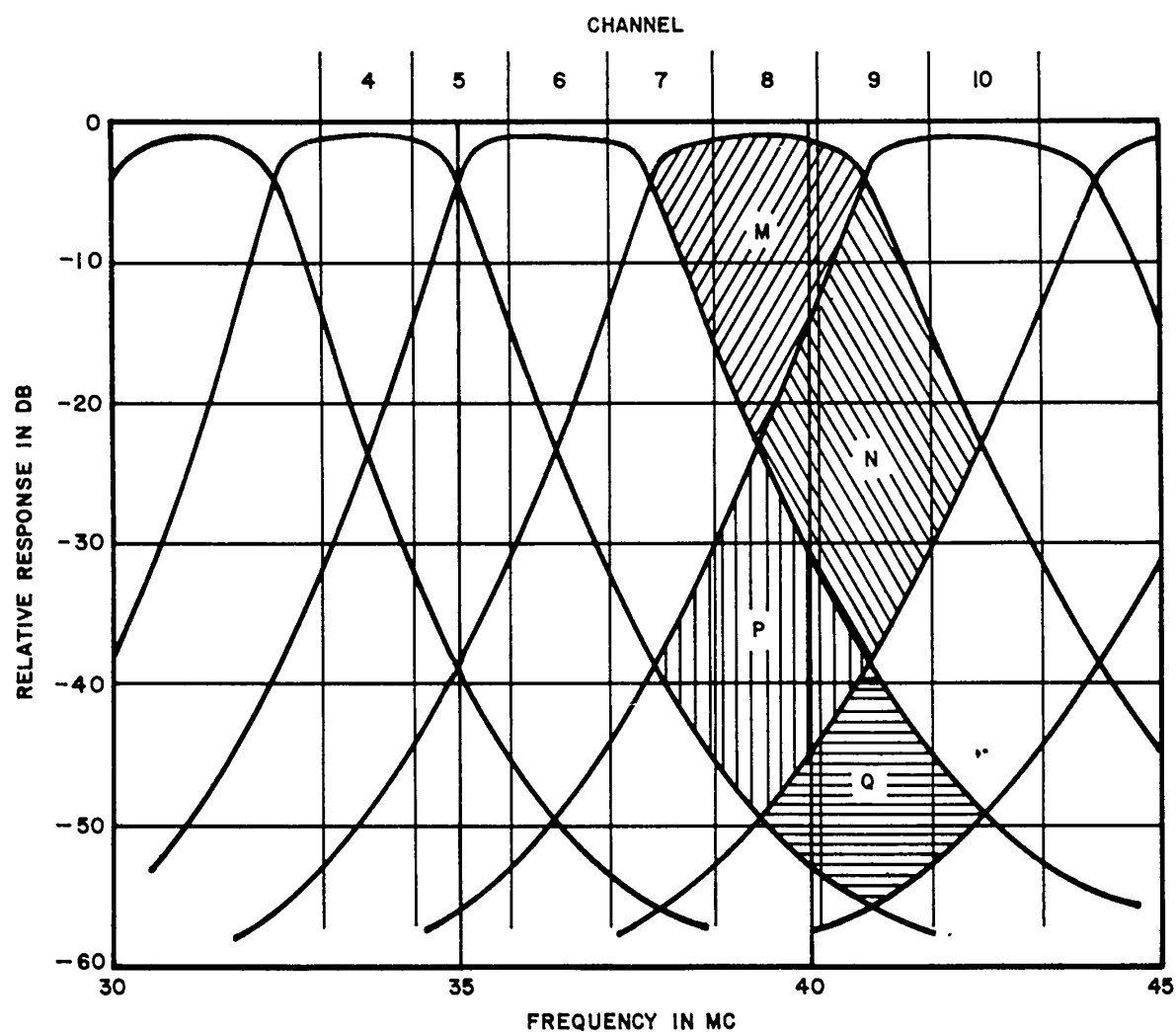
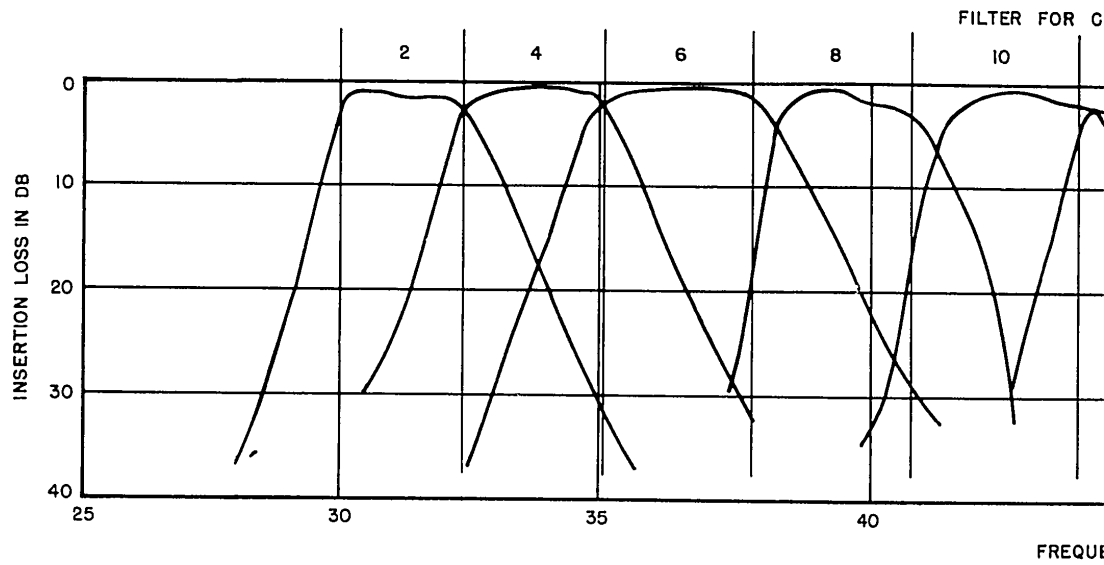


FIGURE 69. REGIONS OF ONE, TWO, THREE, AND FOUR VIDEO-AMPLIFIER RESPONSES

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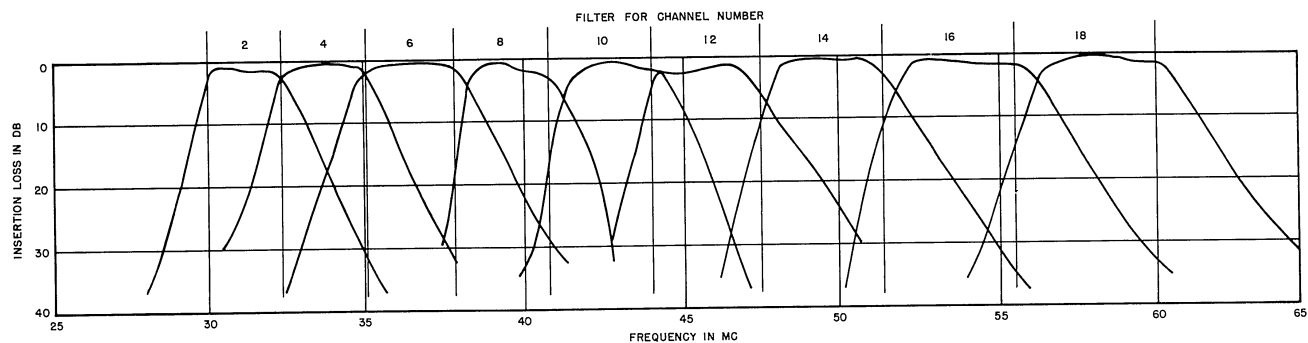


FIGURE 70. INSERTION LOSS VS FREQUENCY OF FILTERS FOR SUB-BAND I

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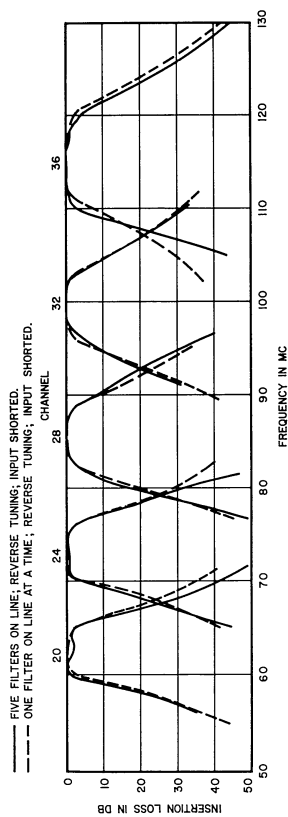


FIGURE 71. INSERTION LOSS VS FREQUENCY OF FILTERS FOR SUB-BAND 2

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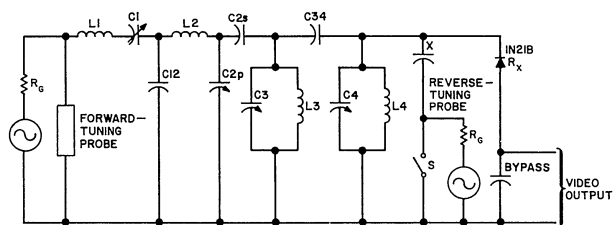


FIGURE 72. FORWARD- AND REVERSE-TUNING PROCEDURES

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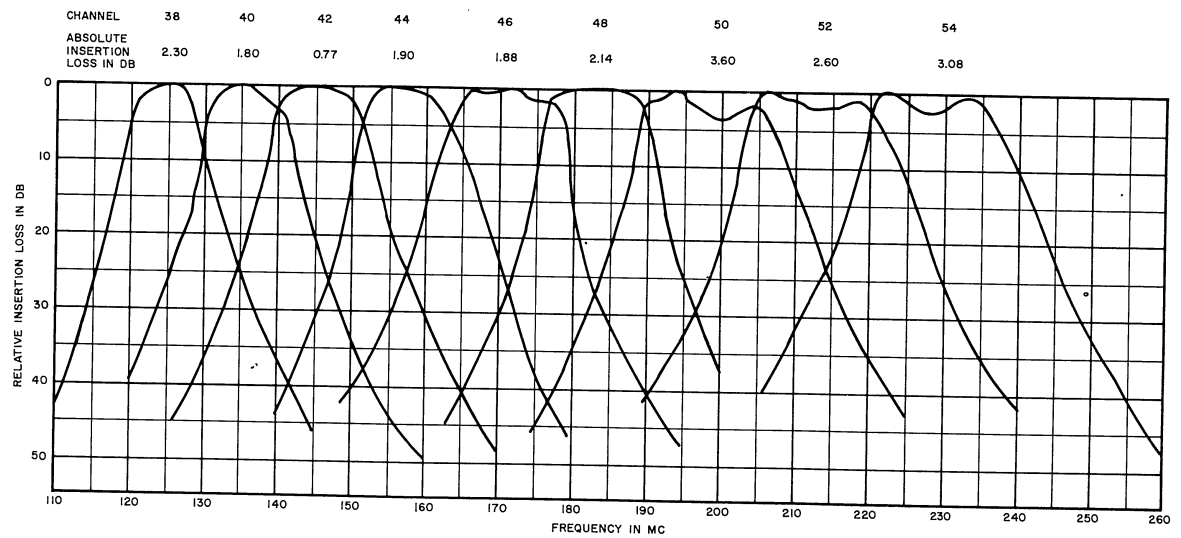


FIGURE 73. INSERTION LOSS VS FREQUENCY OF FILTERS FOR SUB-BAND 3

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FIGURE 73
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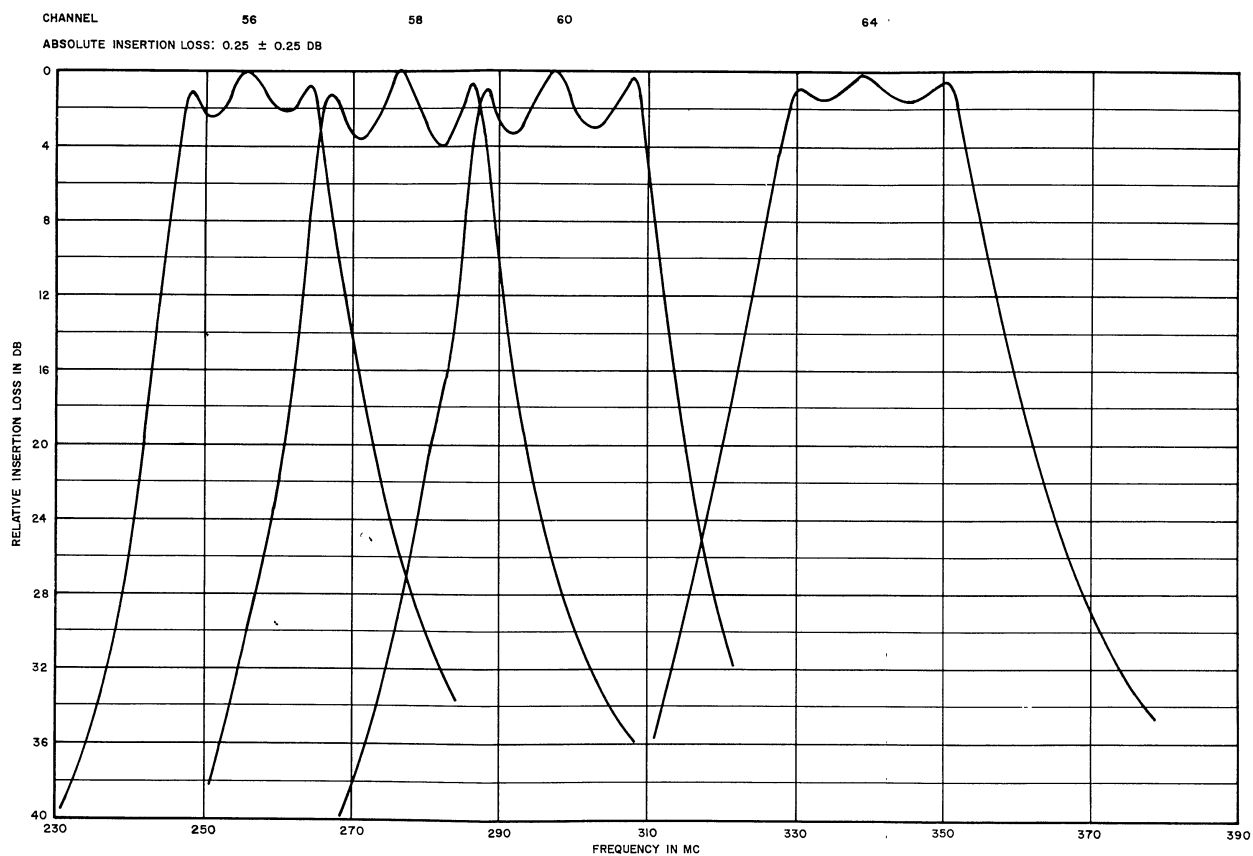


FIGURE 74. INSERTION LOSS VS FREQUENCY OF FILTERS FOR SUB-BAND 4

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FIGURE 74
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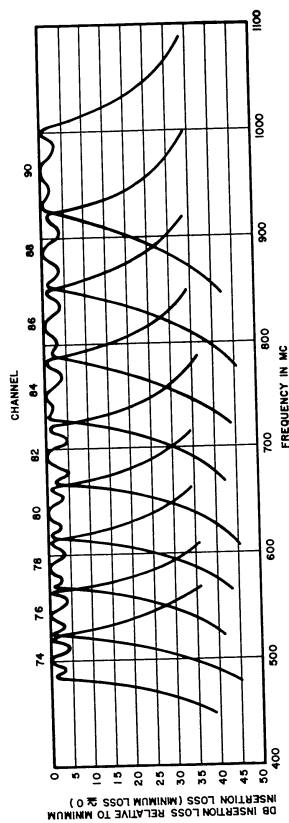


FIGURE 75. INSERTION LOSS VS FREQUENCY OF FILTERS FOR SUB-BAND 5

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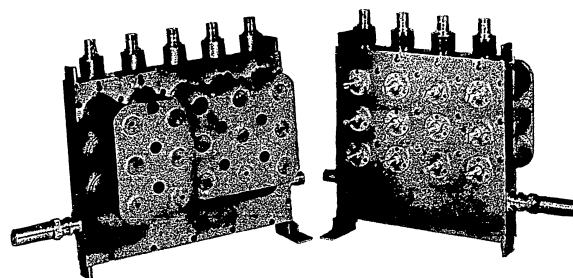


FIGURE 76. ASSEMBLED FILTER GROUPS FOR SUB-BAND 5

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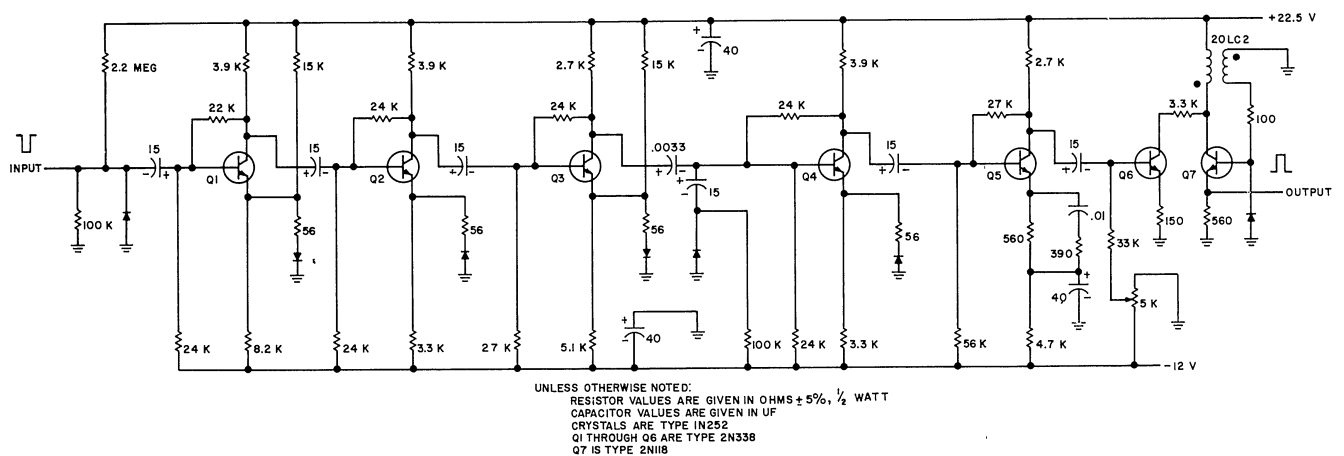


FIGURE 77. SCHEMATIC DIAGRAM OF VIDEO AMPLIFIER FOR FREQUENCY ANALYZER

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FIGURE 77
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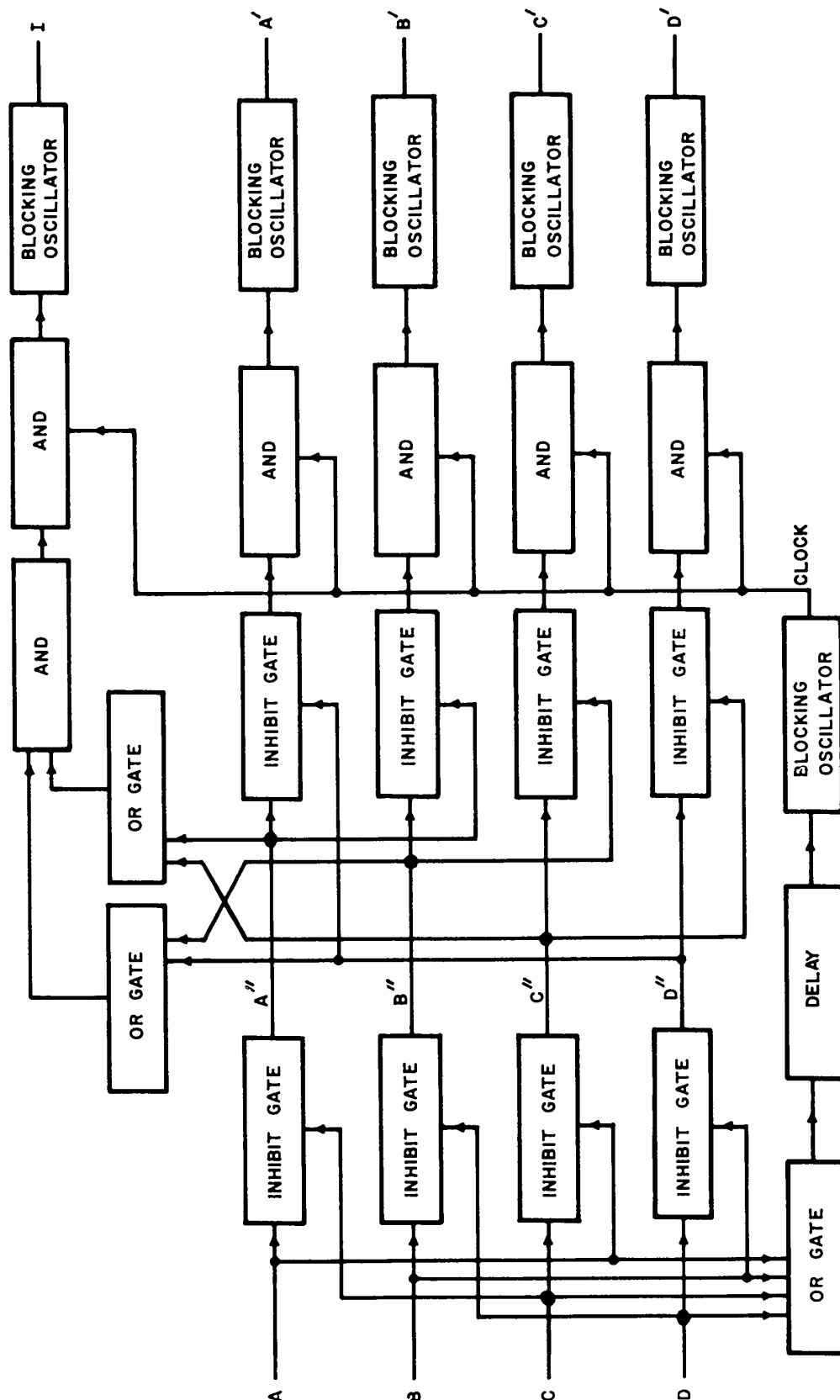


FIGURE 78. BLOCK DIAGRAM OF FREQUENCY-RECEIVER AMEEL

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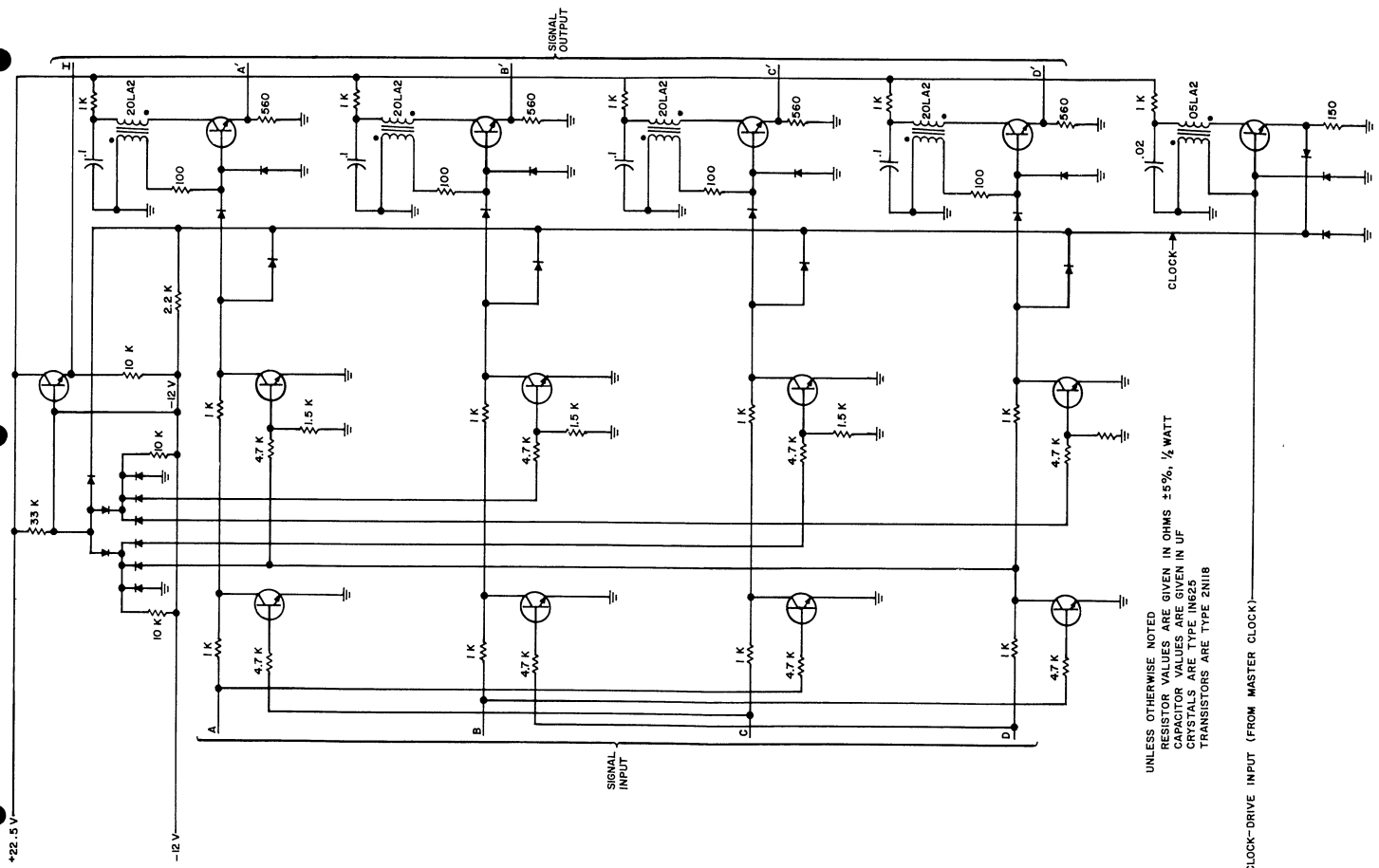


FIGURE 79. SCHEMATIC DIAGRAM OF FREQUENCY-RECEIVER AMBEL

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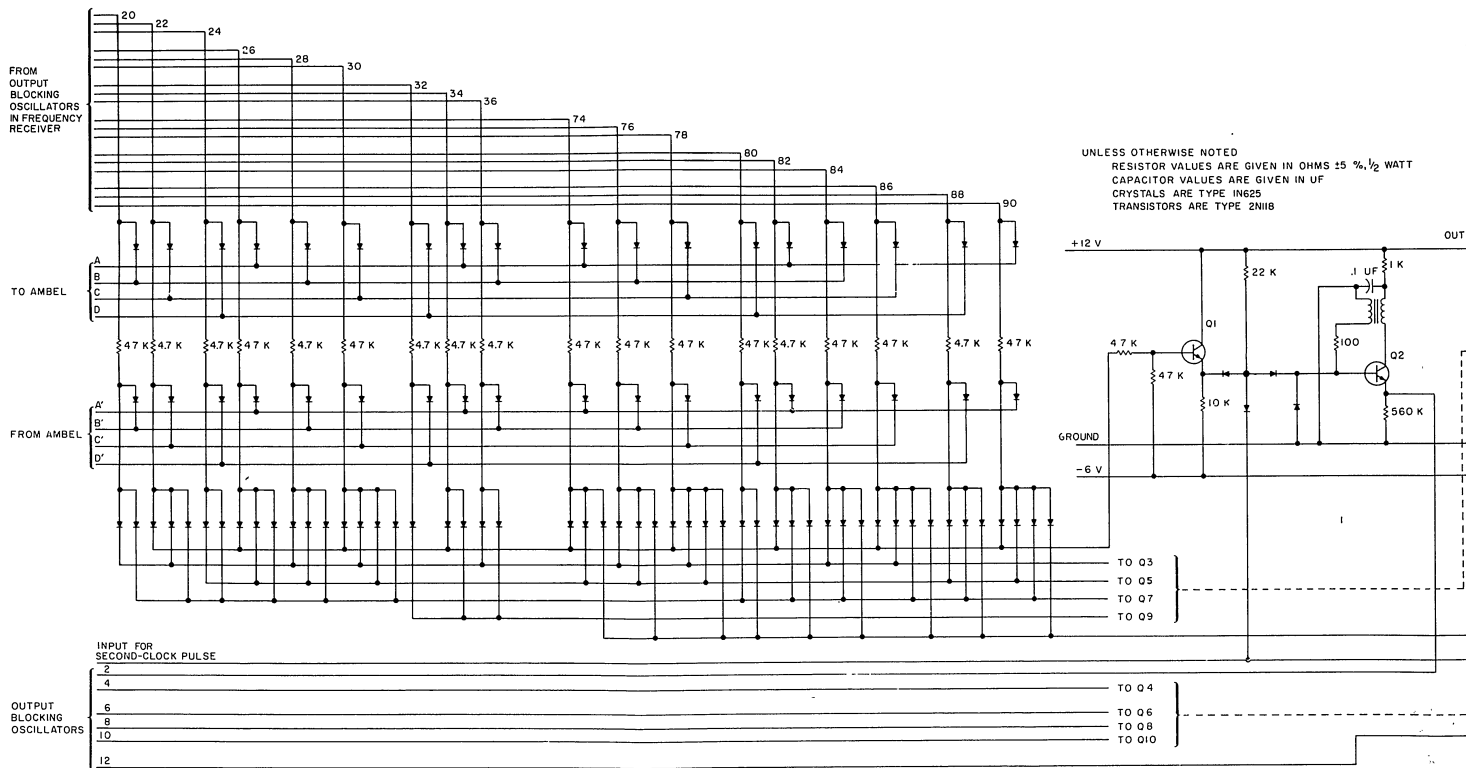
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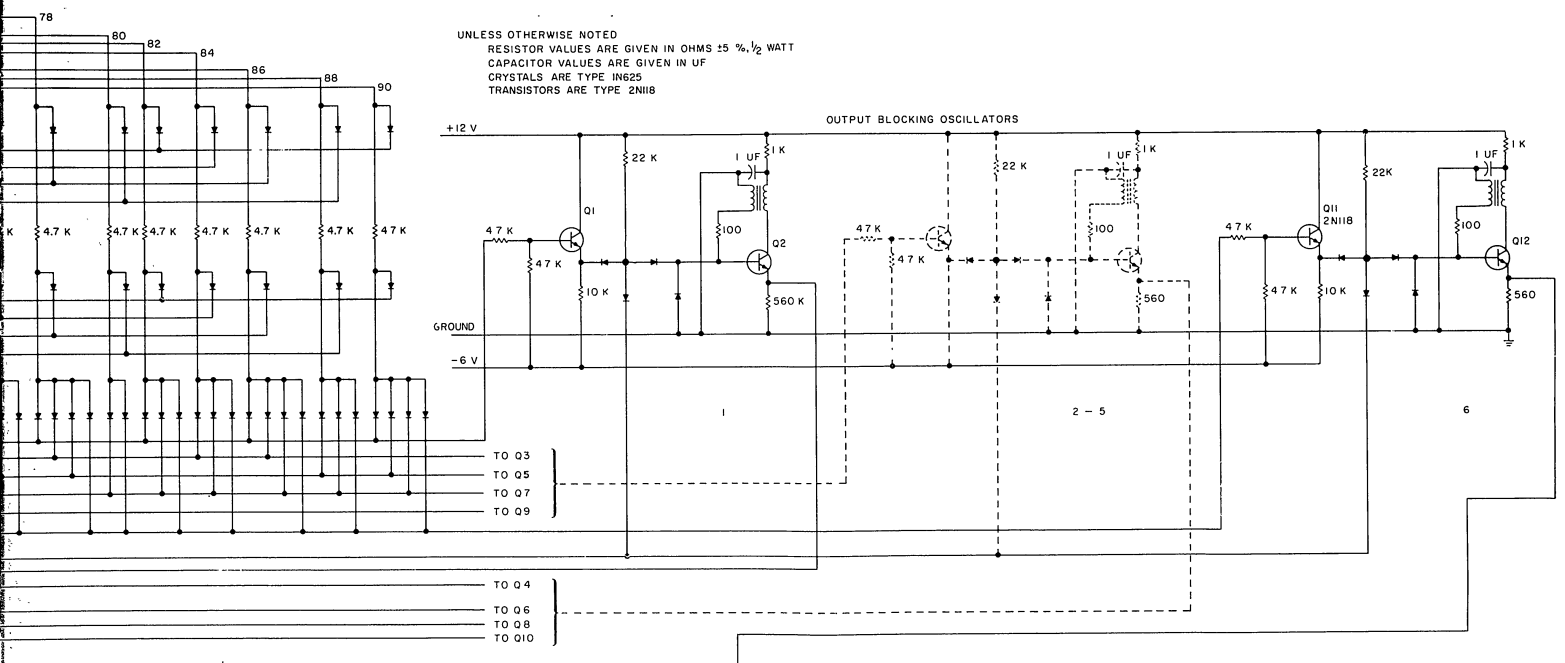


FIGURE 80 SCHEMATIC DIAGRAM OF FREQUENCY-CODING MATRIX FOR SUB-BANDS 2 AND 5

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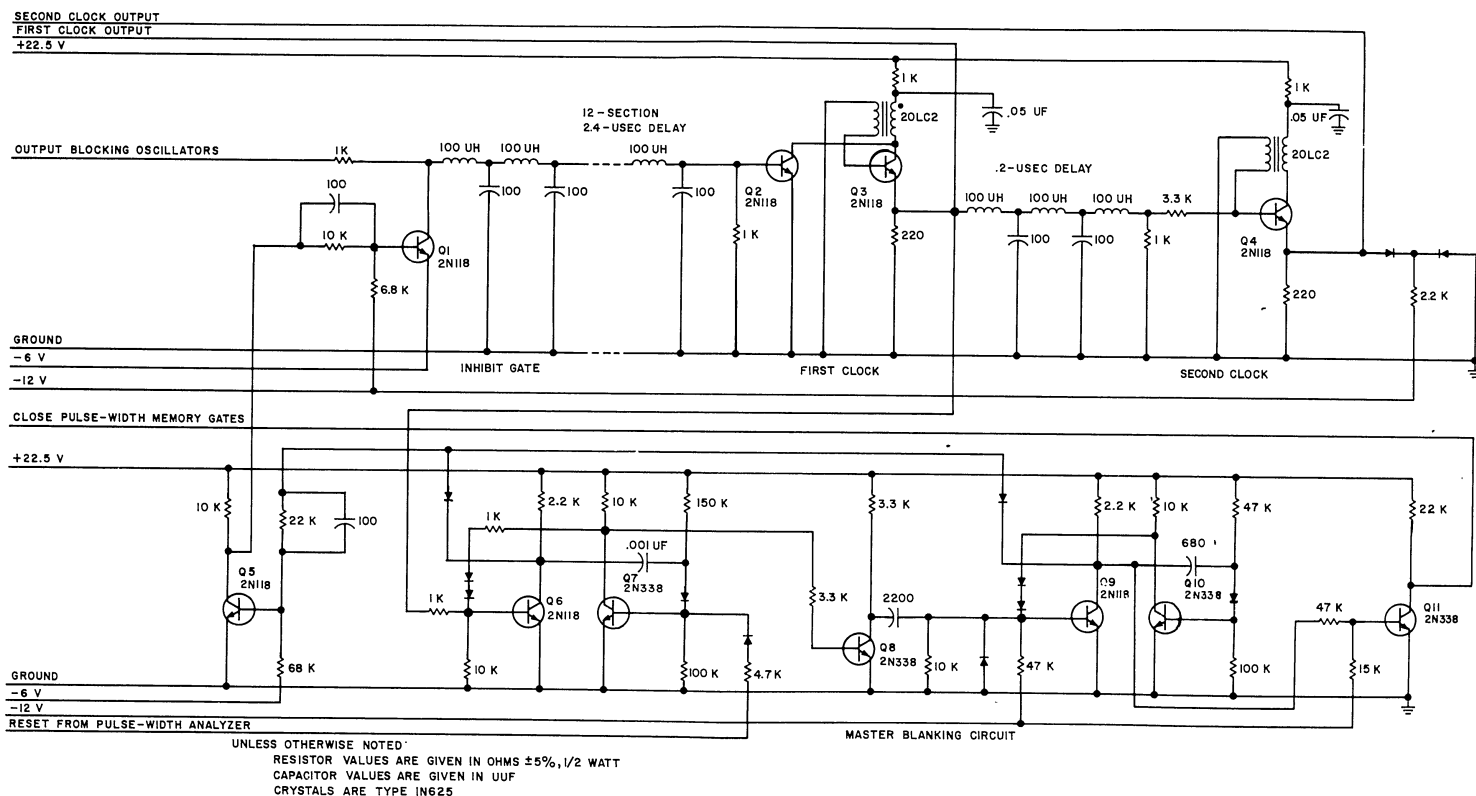


FIGURE 81. SCHEMATIC DIAGRAM OF CENTRAL CLOCK

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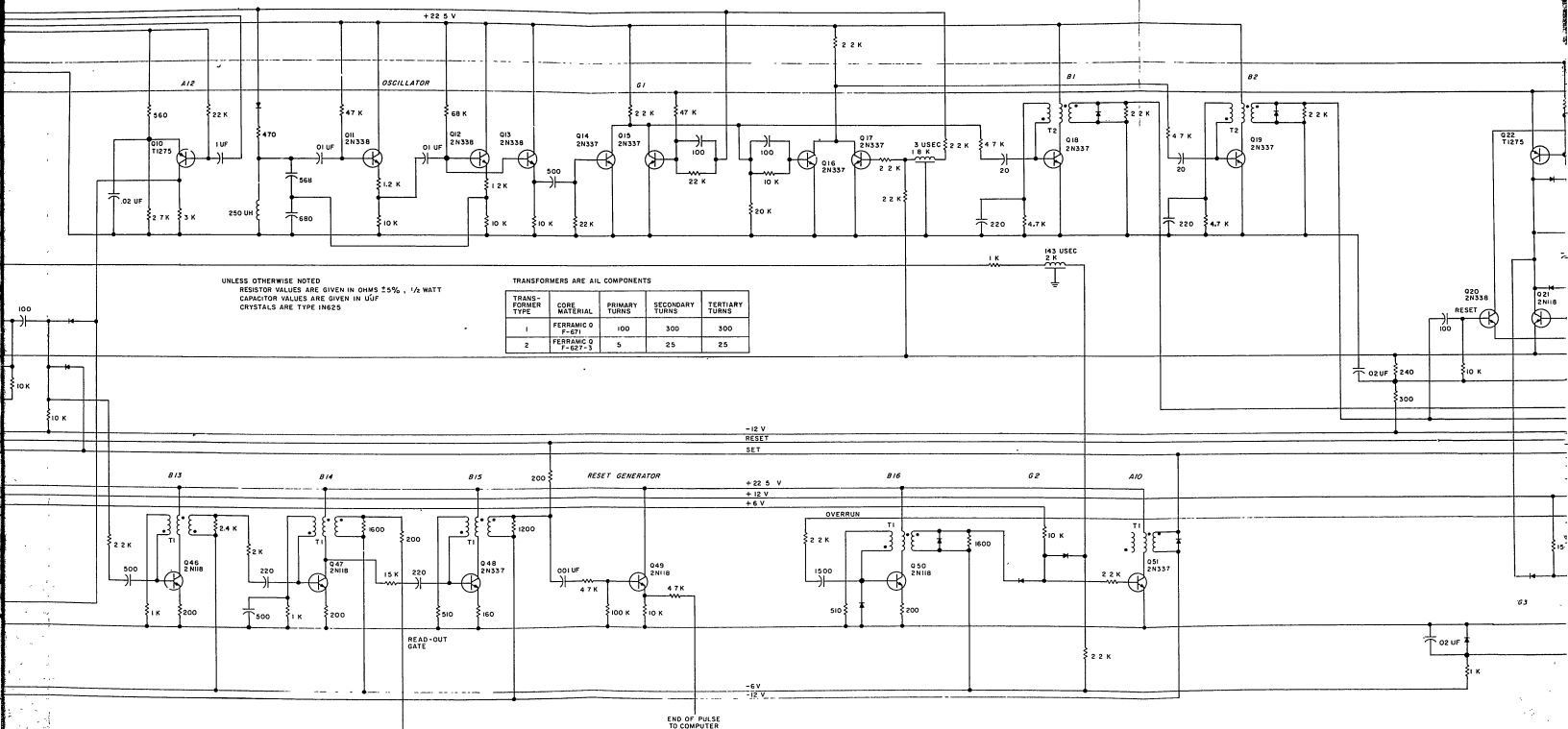




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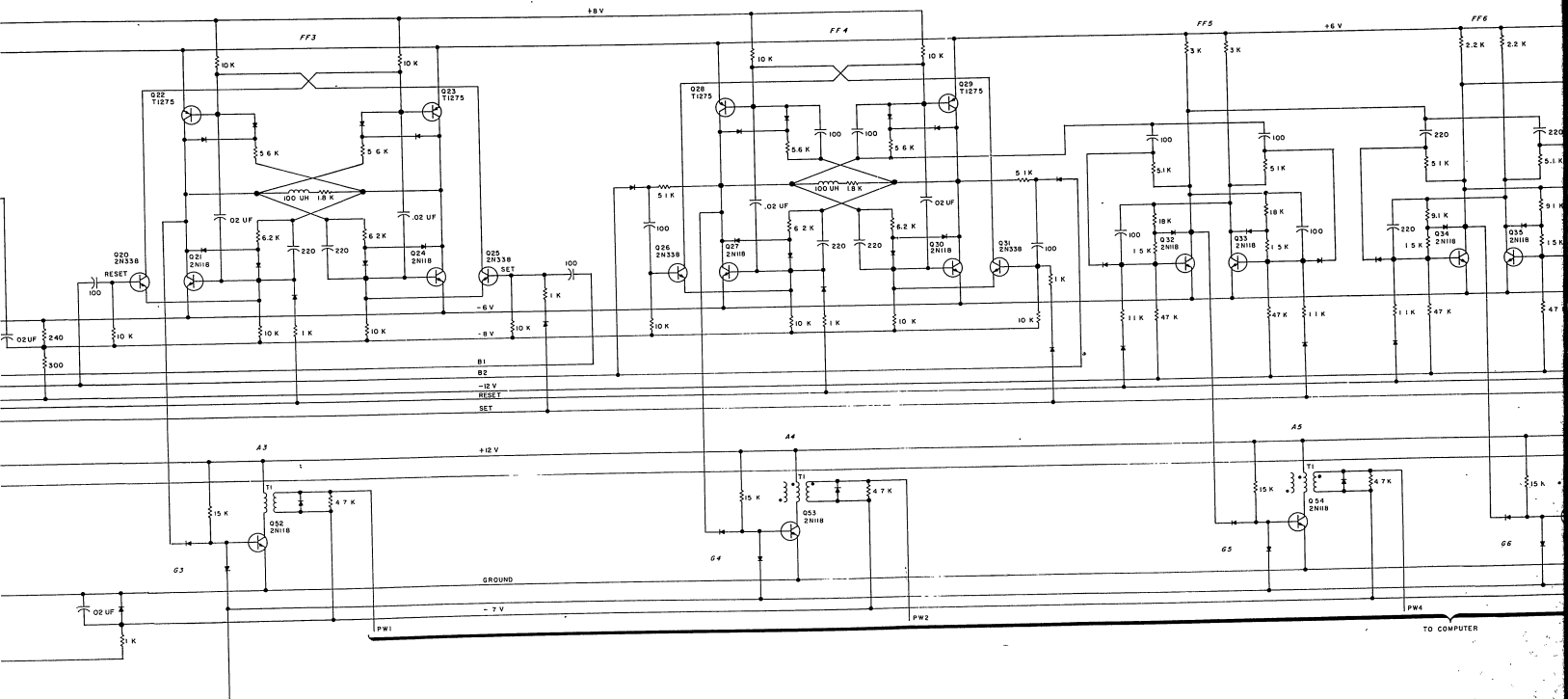
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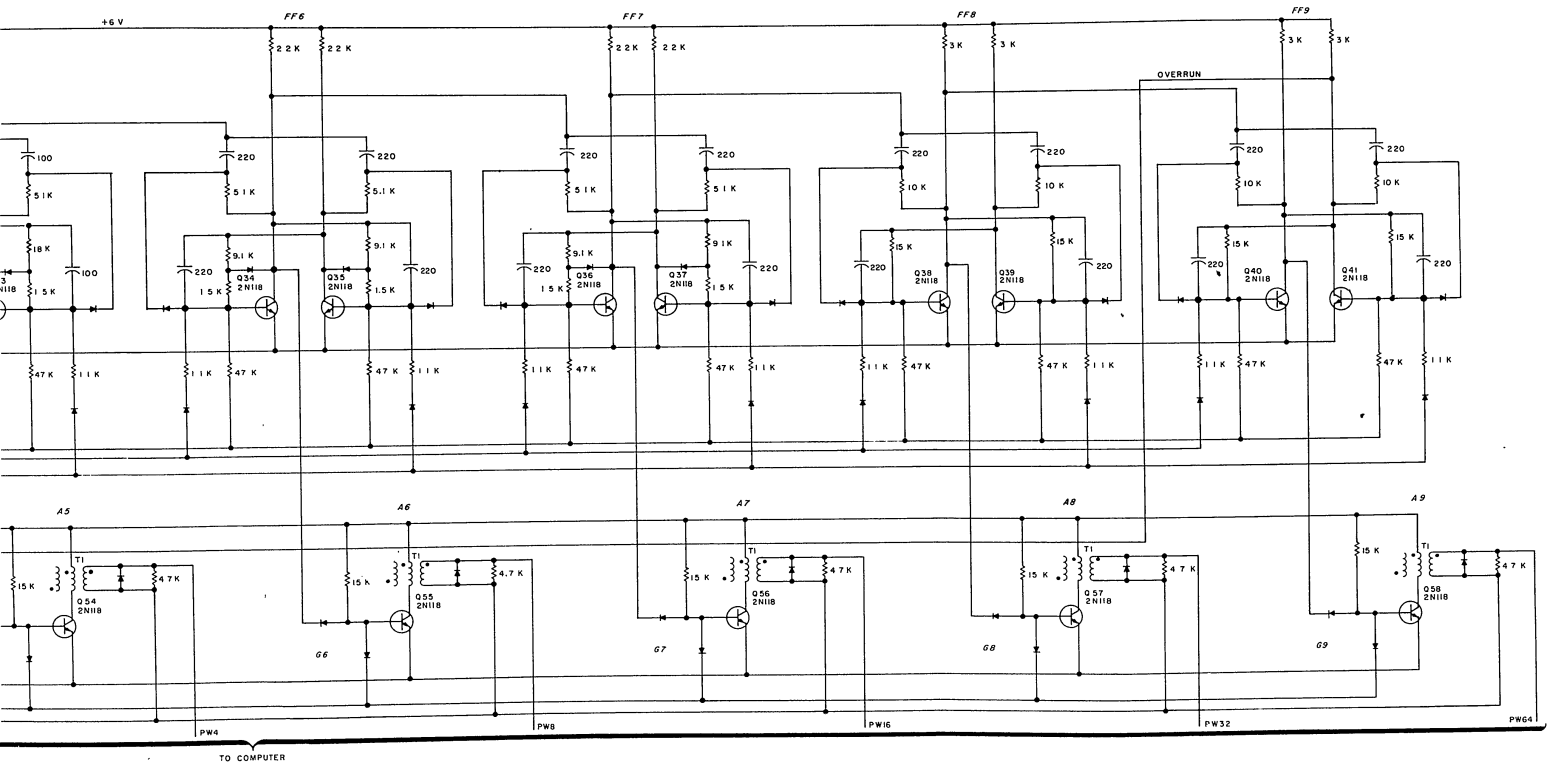


FIGURE 83. SCHEMATIC DIAGRAM OF PULSE-WIDTH ANALYZER

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FIGURE 83
297

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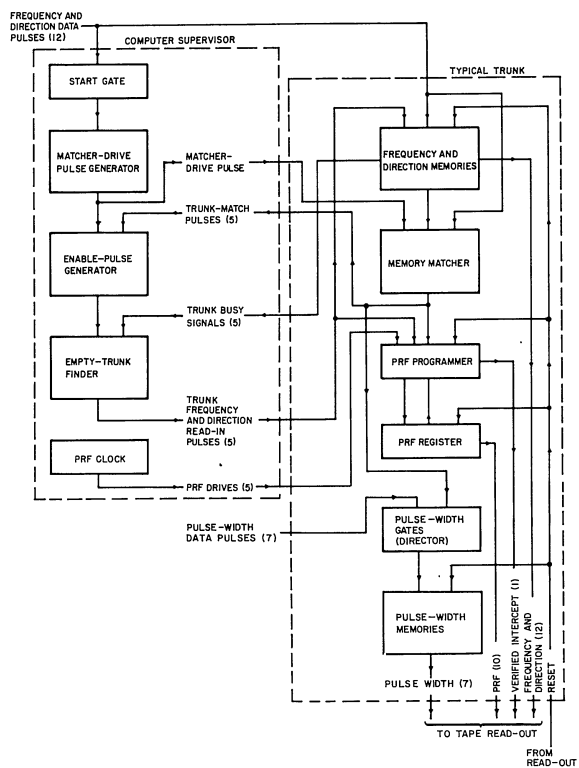


FIGURE 84. BLOCK DIAGRAM OF COMPUTER

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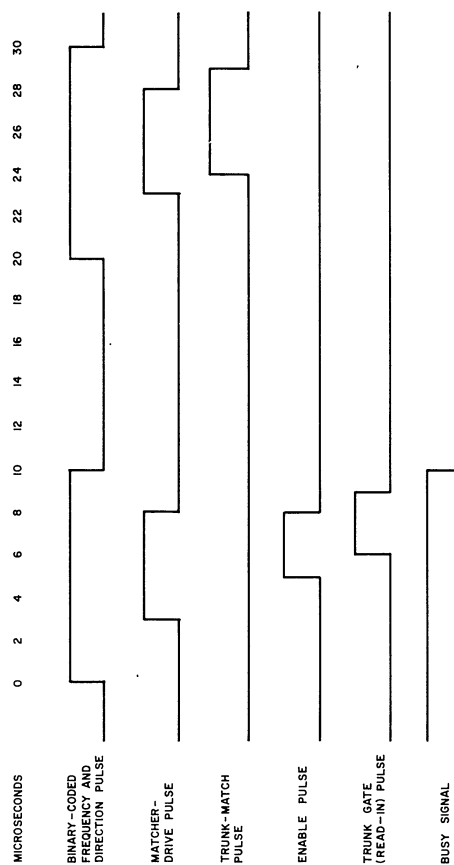


FIGURE 85. TIMING DIAGRAM OF COMPUTER

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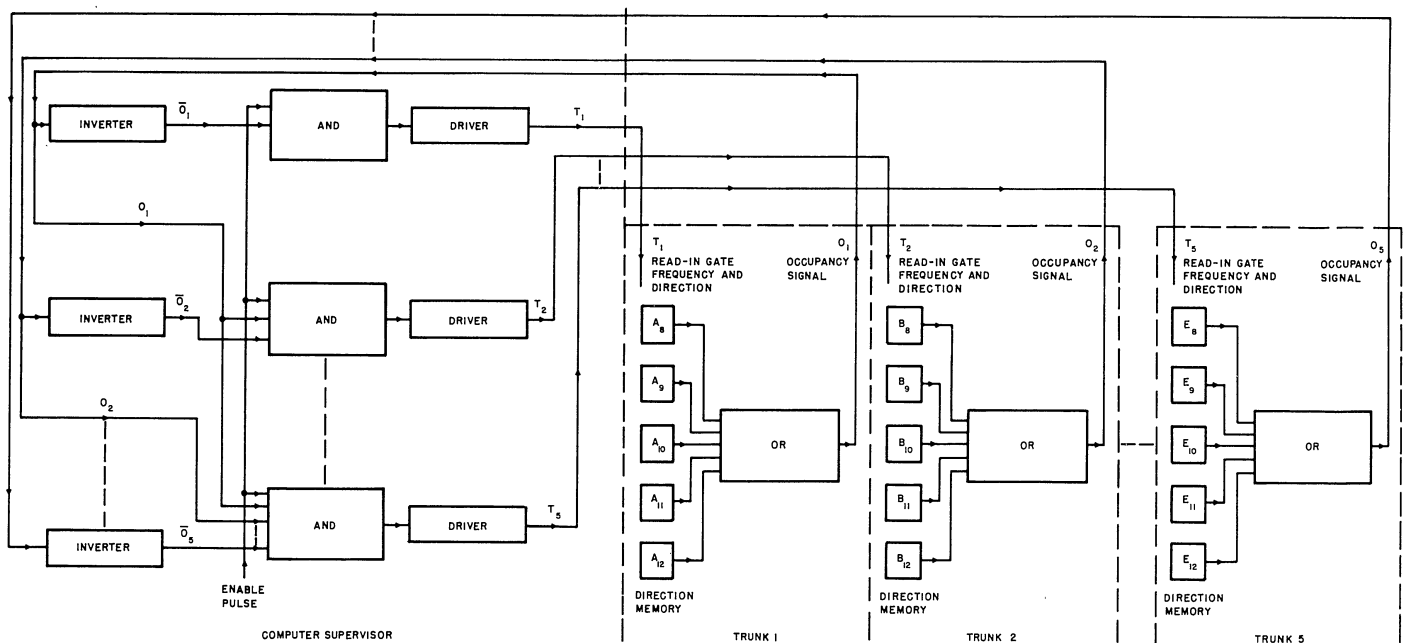


FIGURE 86. BLOCK DIAGRAM OF EMPTY-TRUNK FINDER

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FIGURE 86
303

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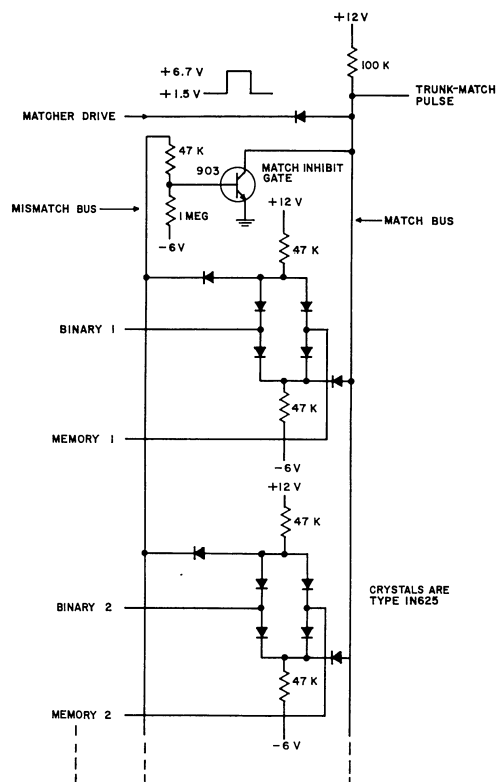


FIGURE 87. SCHEMATIC DIAGRAM OF MATCHER

305

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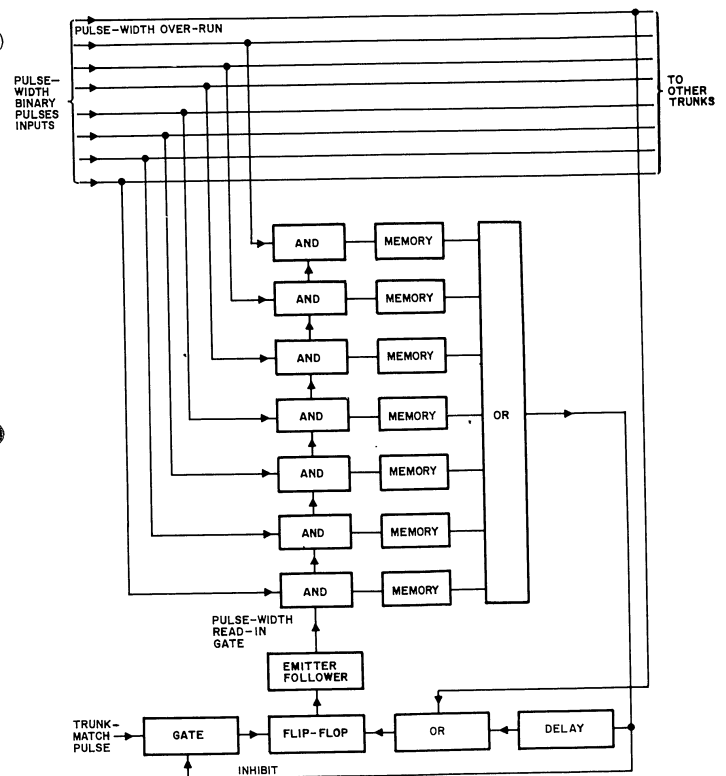


FIGURE 88. BLOCK DIAGRAM OF PULSE-WIDTH DIRECTOR AND MEMORIES

307

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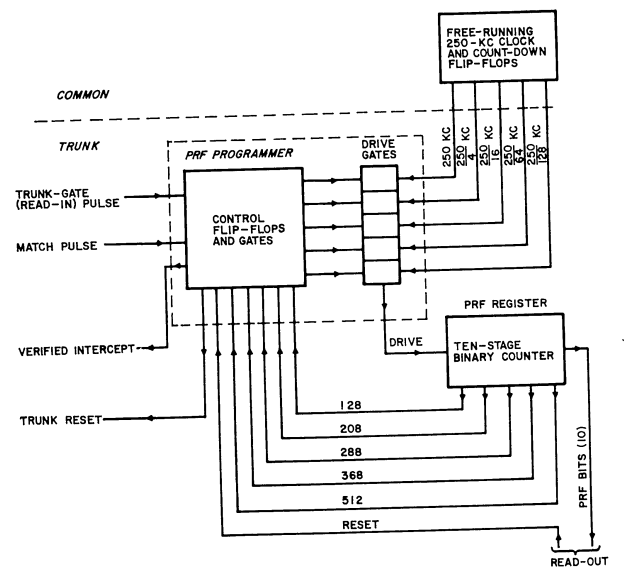


FIGURE 89. SIMPLIFIED BLOCK DIAGRAM OF PRF ANALYZER

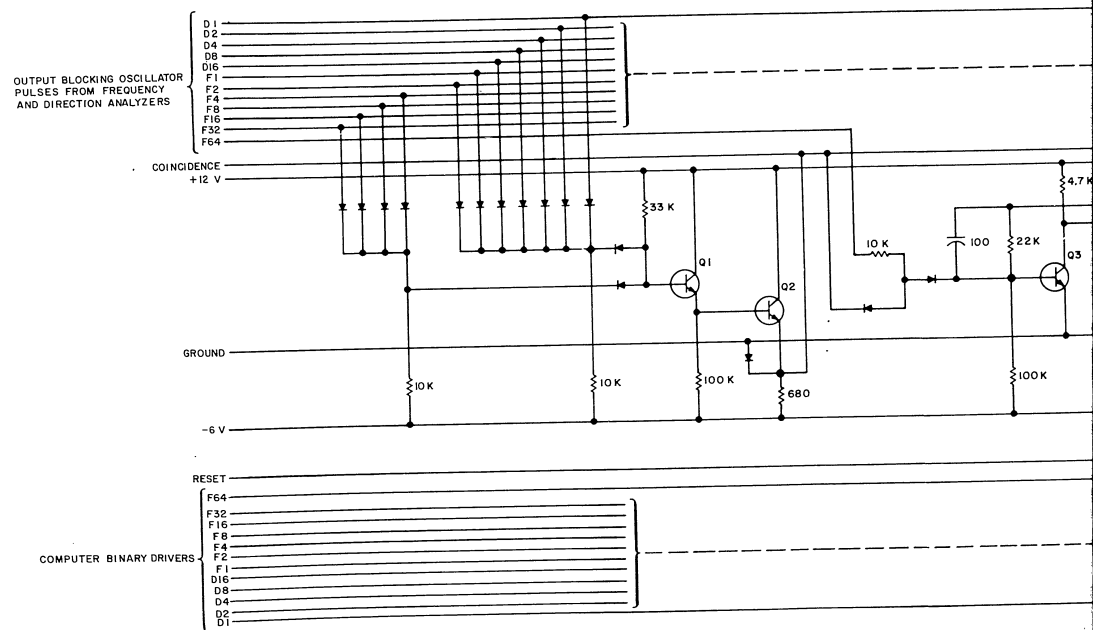
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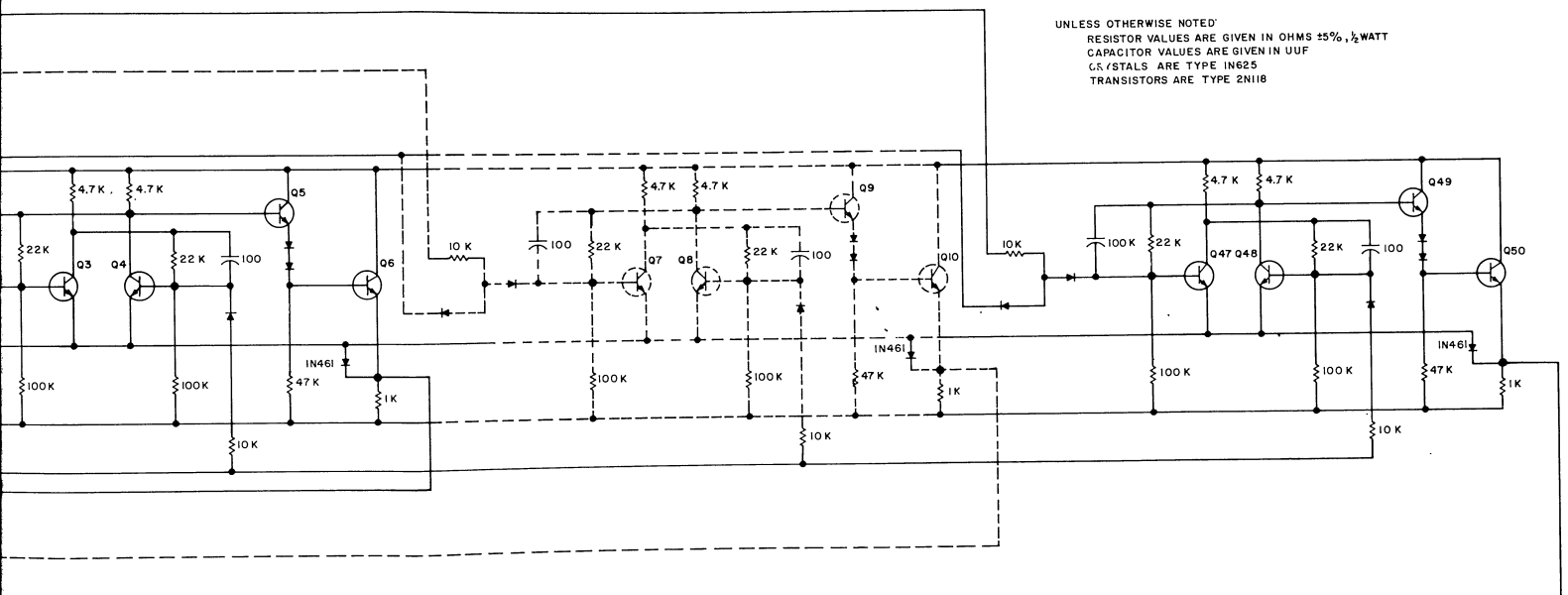
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A. BINARY DRIVER

FIGURE 90. SCHEMATIC DIAGRAM OF COMPUTER SUPERVISOR

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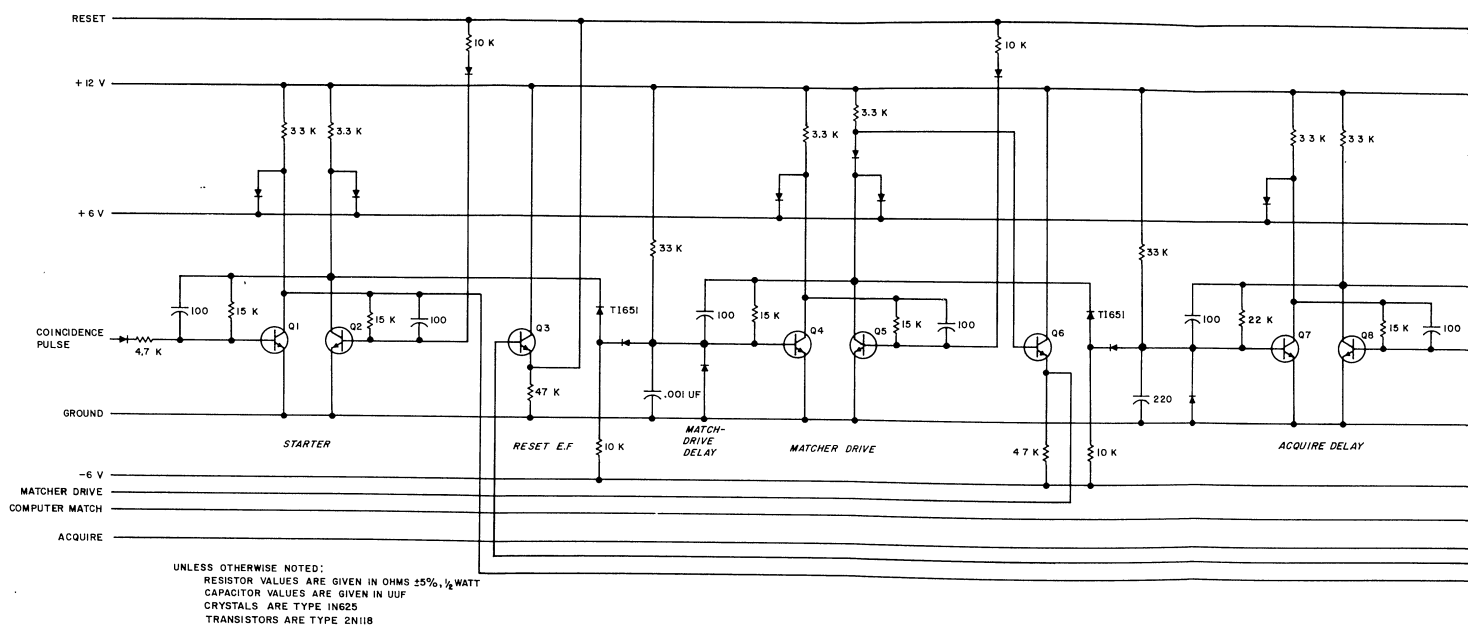
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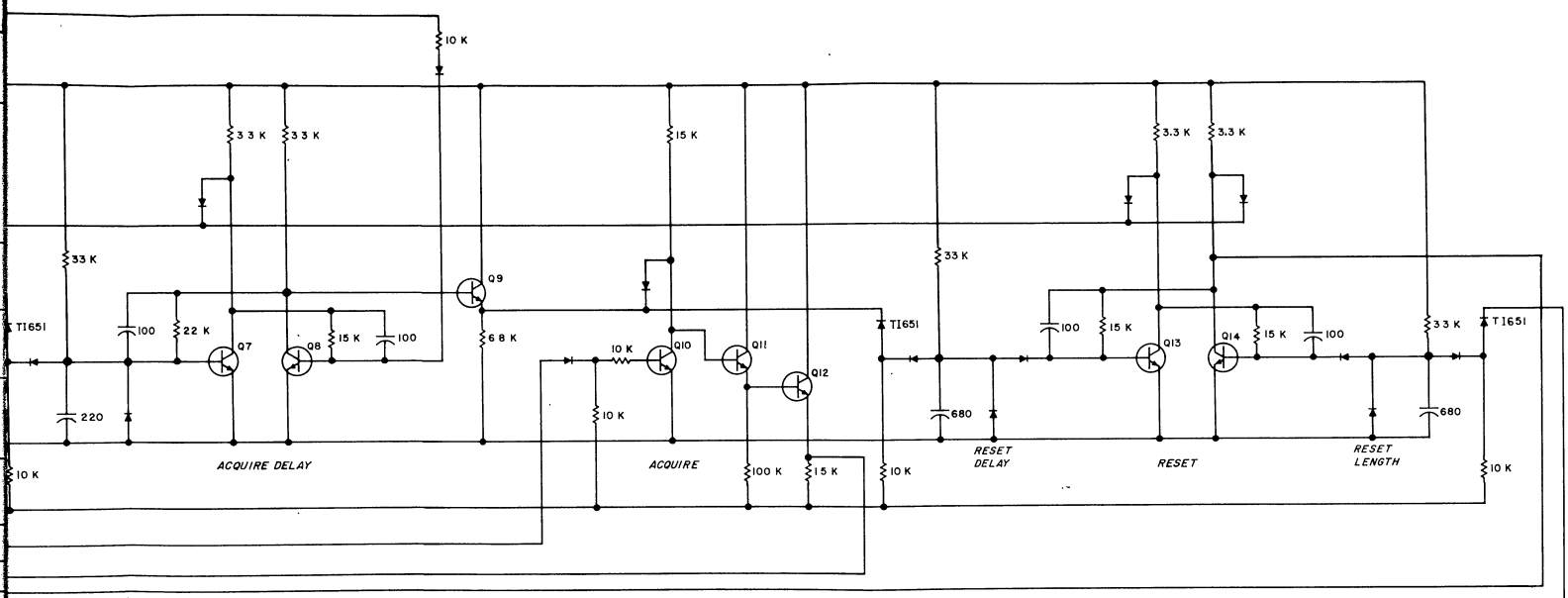
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B. PROGRAMMER

FIGURE 90

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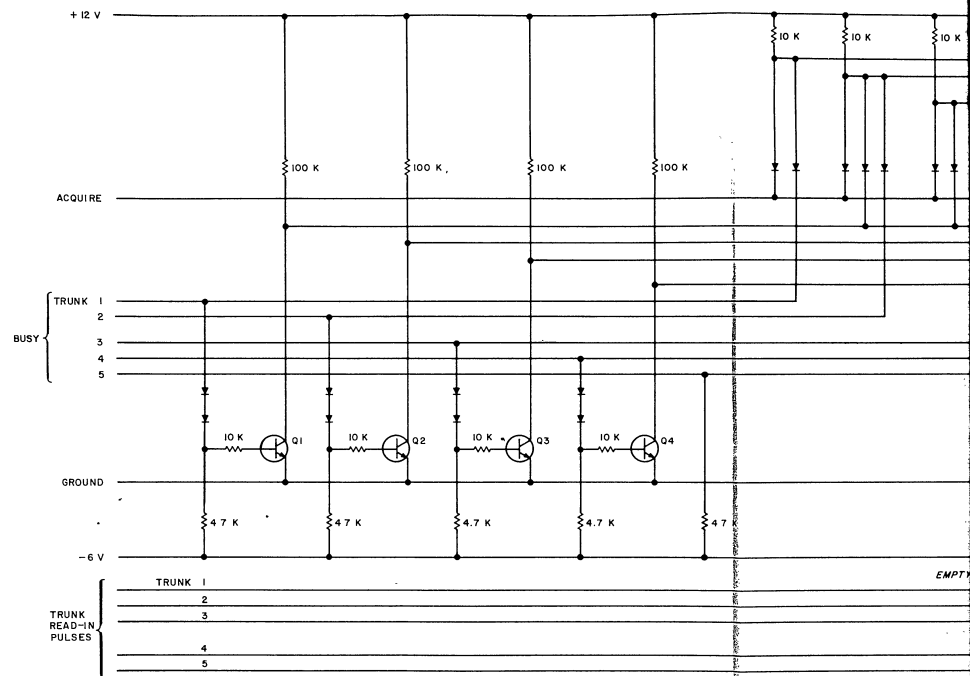
SHEET 2 OF 4
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UNLESS OTHERWISE NOTED:
RESISTOR VALUES ARE GIVEN IN OHMS $\pm 5\%$, $\frac{1}{2}$ WATT
CRYSTALS ARE TYPE 1N625
TRANSISTORS ARE TYPE 2N118

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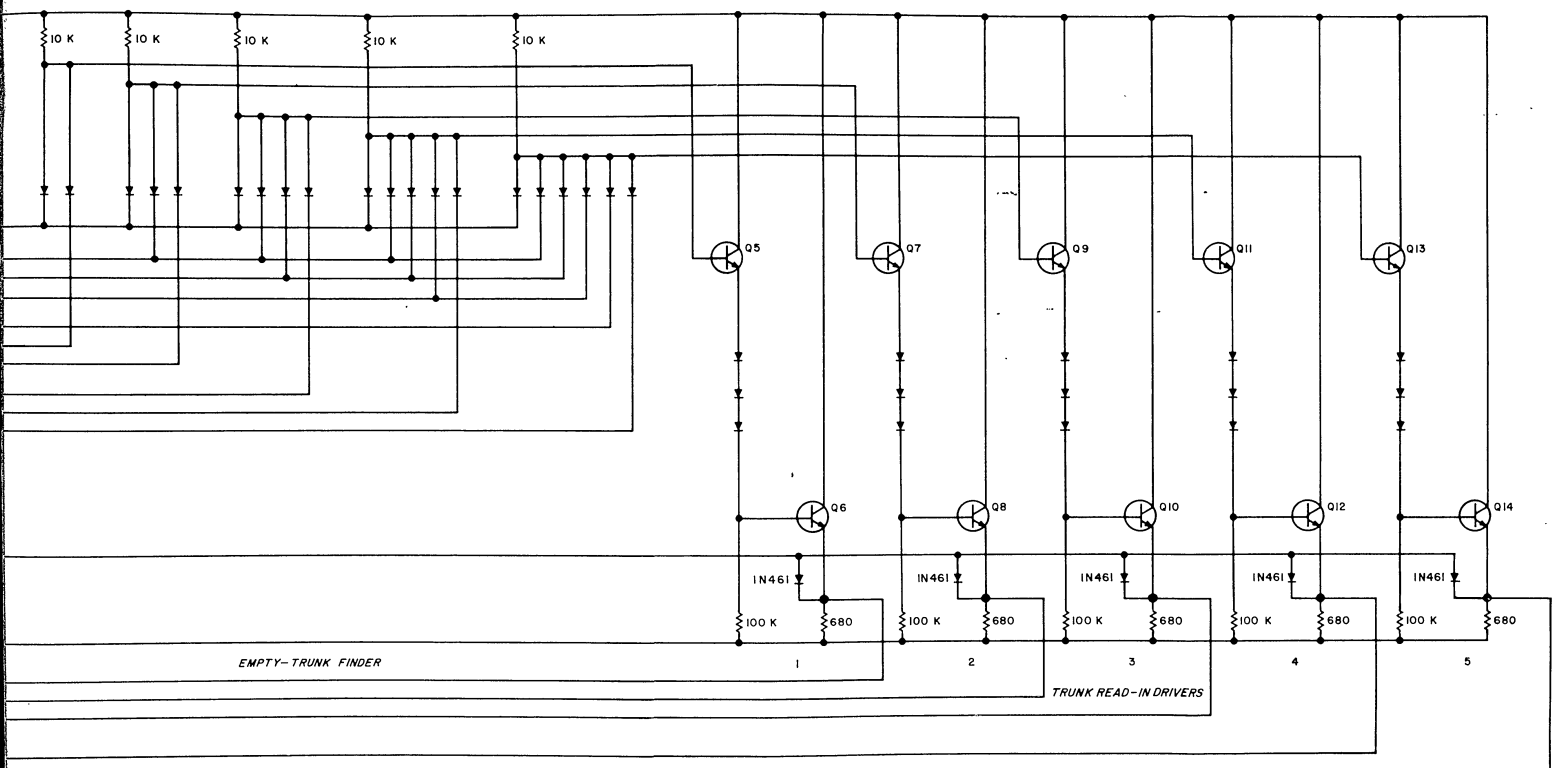
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C. EMPTY-TRUNK FINDER AND READ-IN DRIVERS

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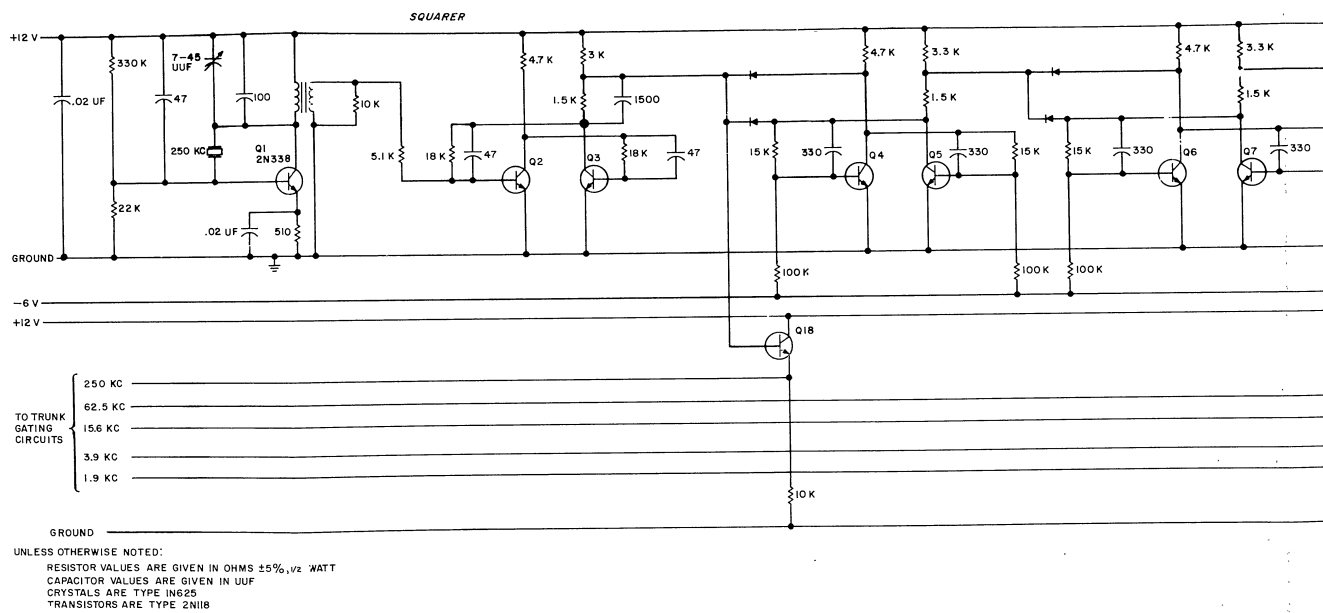
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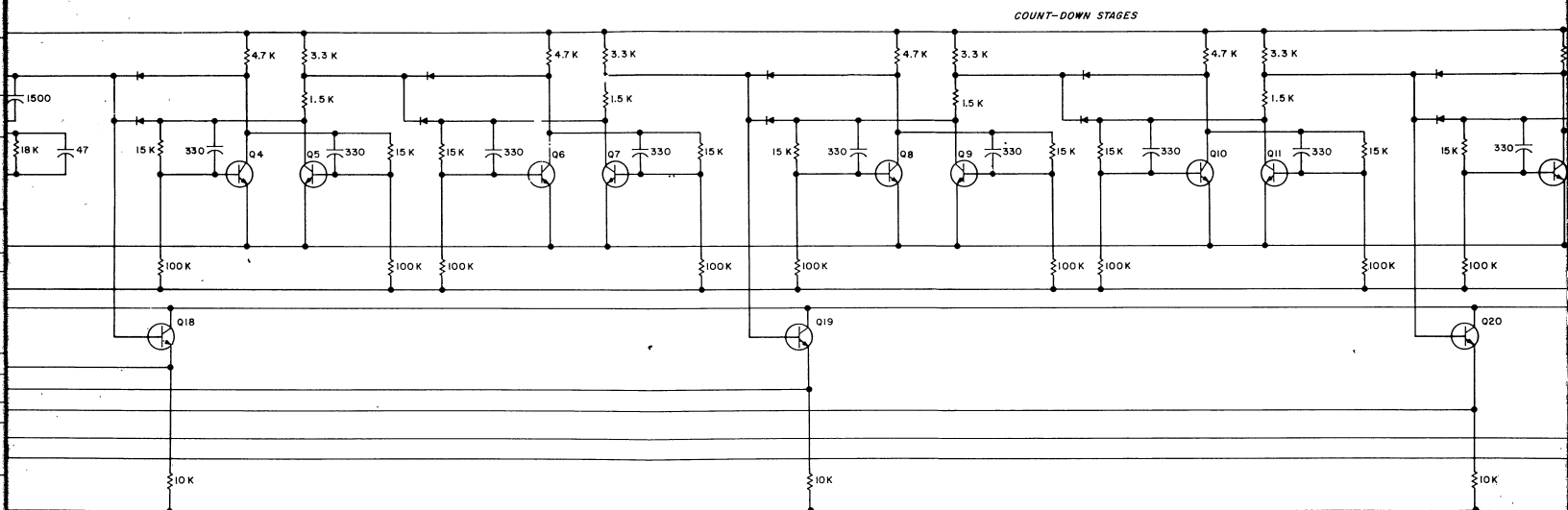
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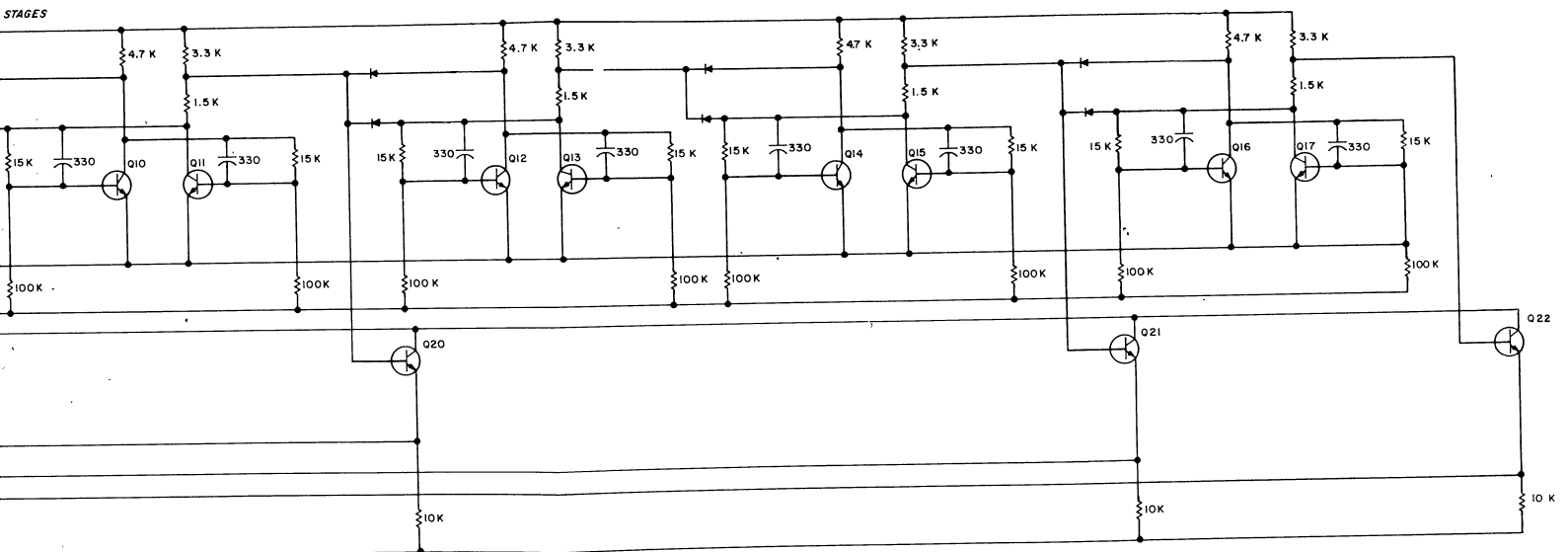
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D. COMMON CLOCKS FOR PRF COUNTER

FIGURE 90

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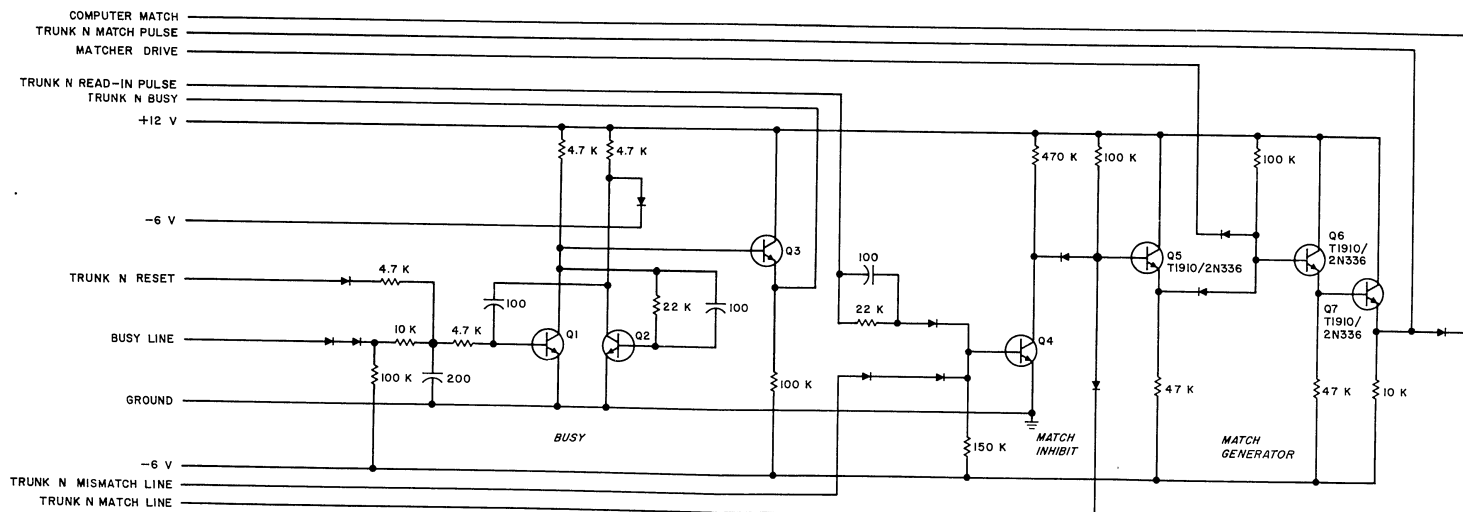
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UNLESS OTHERWISE NOTED:
 RESISTOR VALUES ARE GIVEN IN OHMS $\pm 5\%$, $\frac{1}{2}$ WATT
 CAPACITOR VALUES ARE GIVEN IN UUF
 CRYSTALS ARE TYPE IN625
 TRANSISTORS ARE TYPE 2N118

A. BUSY AND MATCHER CIRCUITS

FIGURE 91. SCHEMATIC DIAGRAM OF COMPUTER TRUNK

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FIGURE 91
SHEET 1 OF 4
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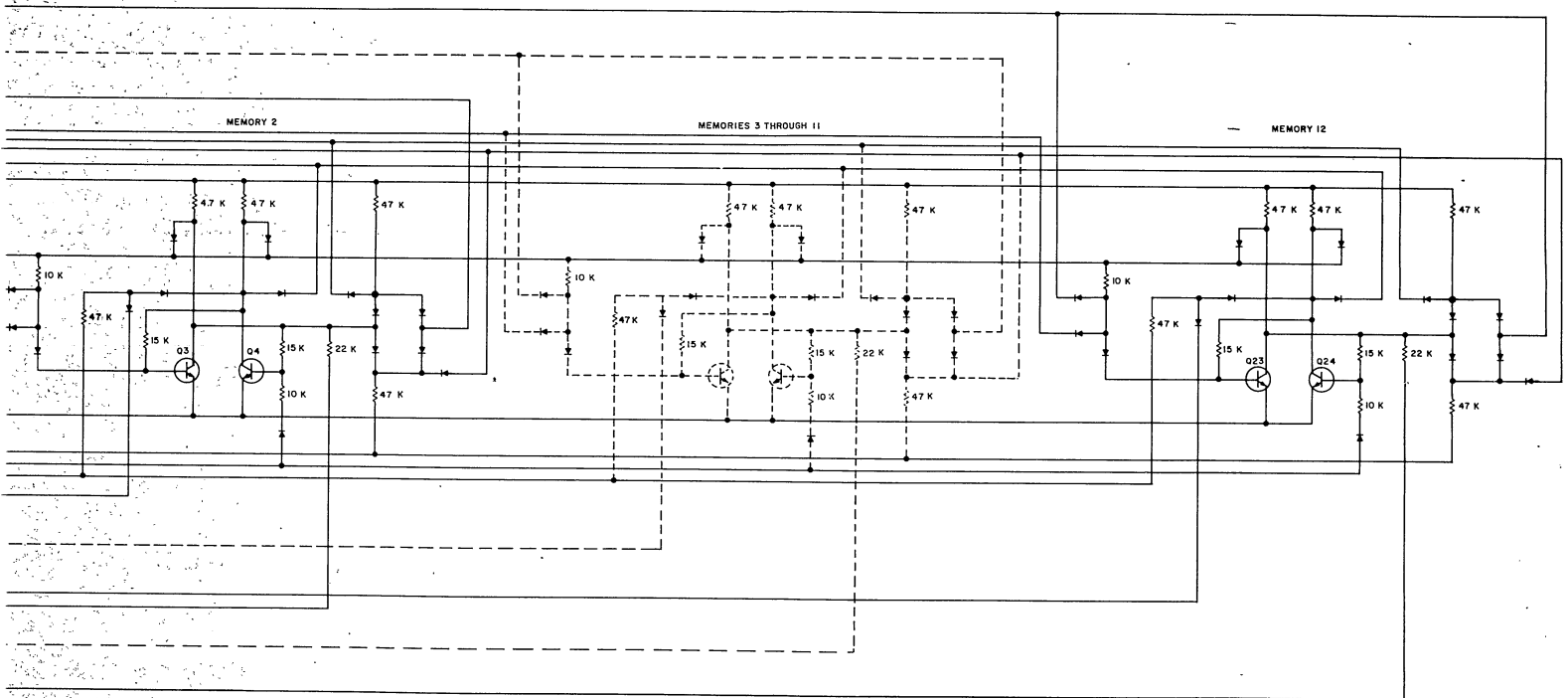


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B. FREQUENCY AND DIRECTION MEMORIES

FIGURE 9L

SHEET 2 OF 4
321

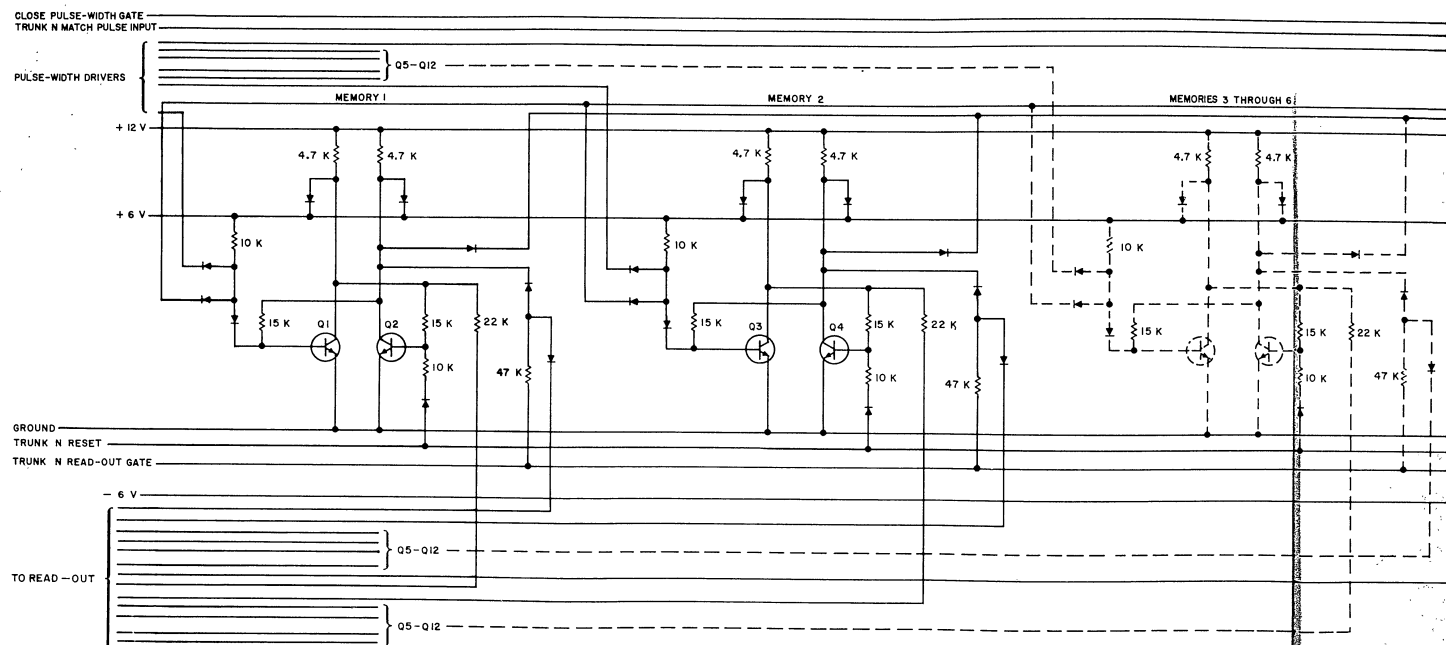
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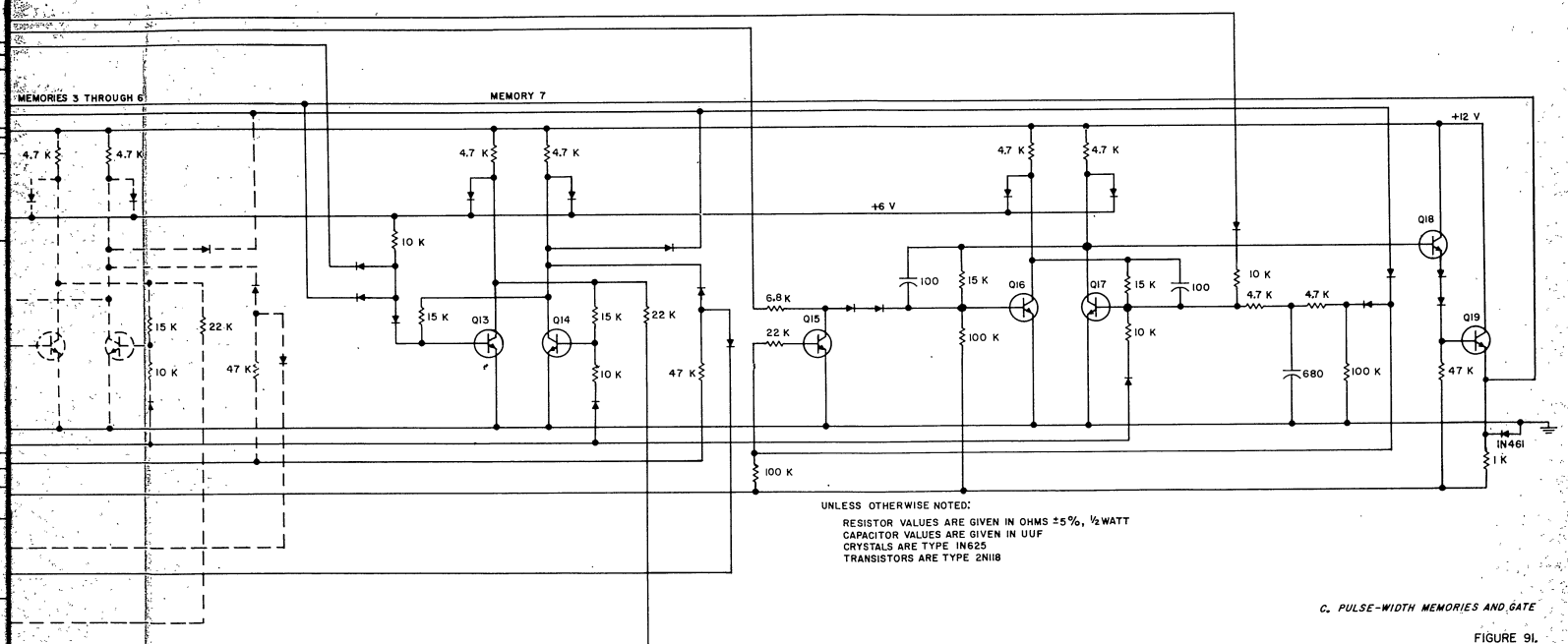
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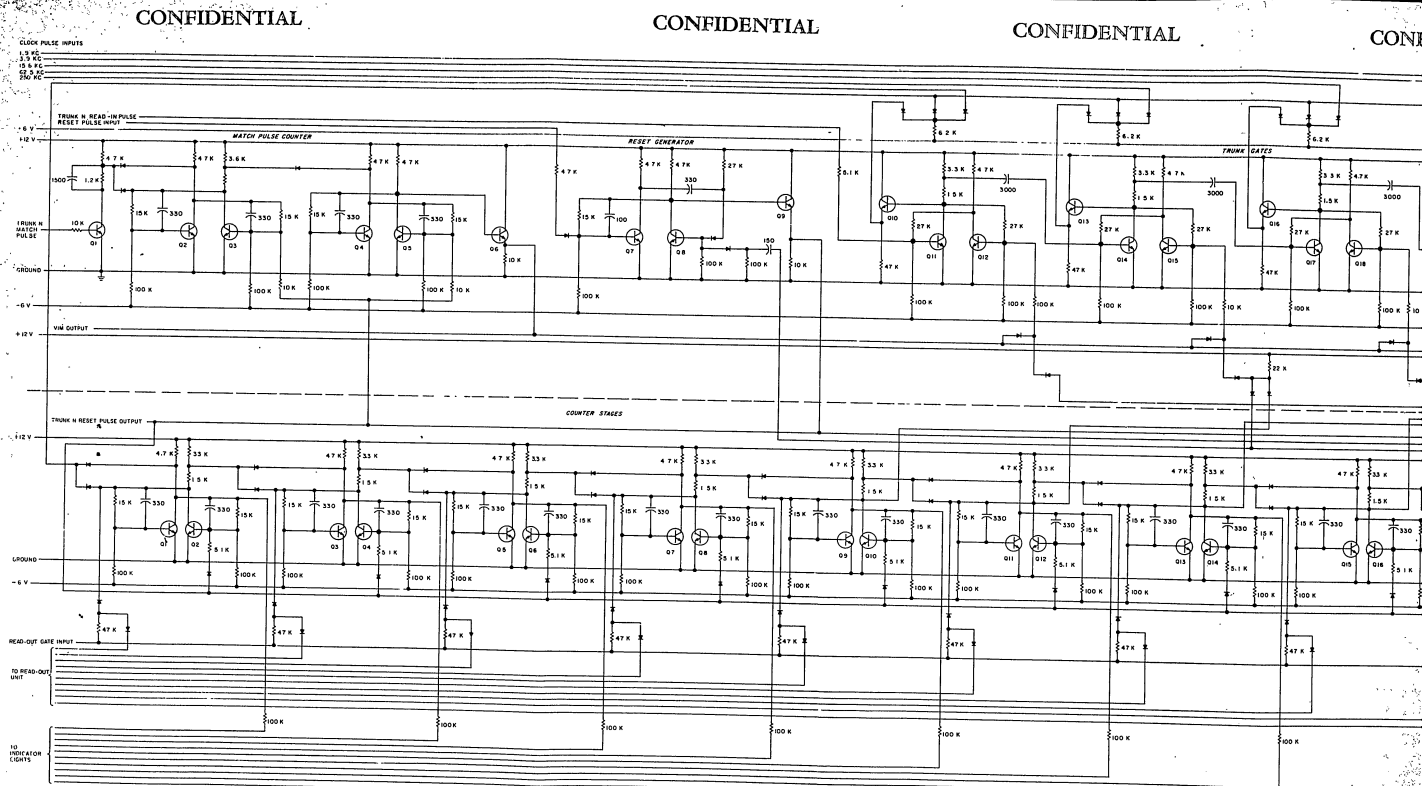
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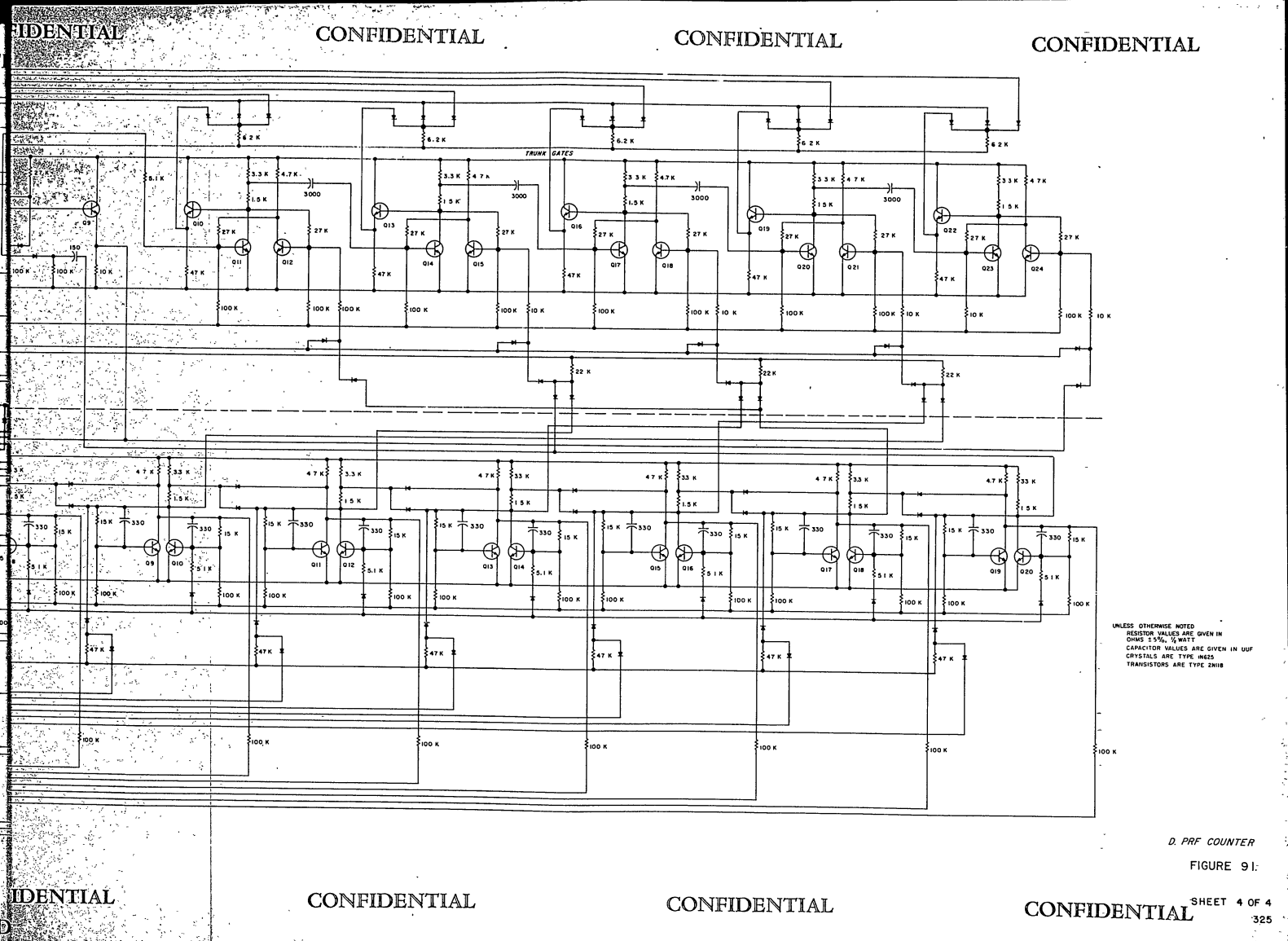


C. PULSE-WIDTH MEMORIES AND GATE

FIGURE 91.

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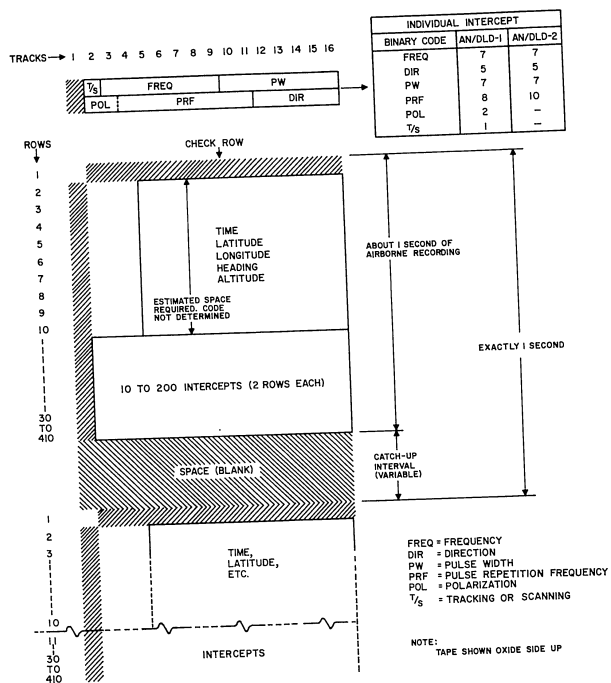


FIGURE 92. RECOMMENDED TAPE FORMAT FOR AN/DLD-1 AND AN/DLD-2

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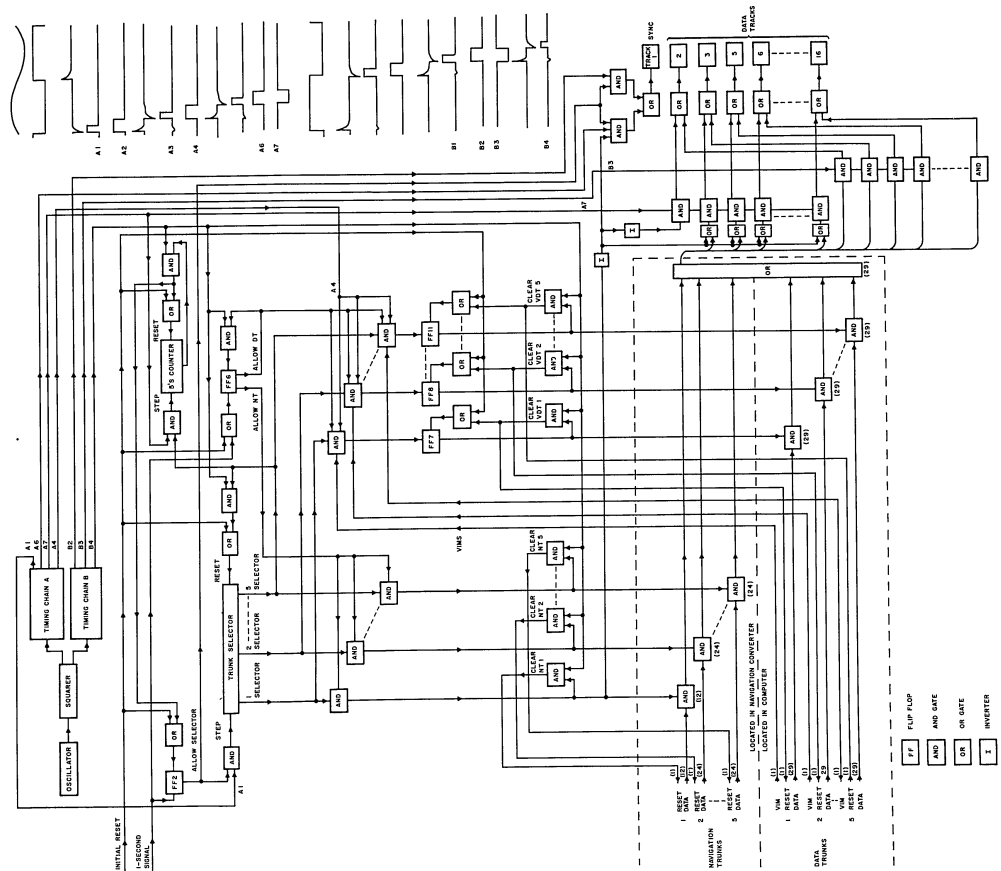


FIGURE 93. TIMING AND BLOCK DIAGRAM OF READ-OUT UNIT

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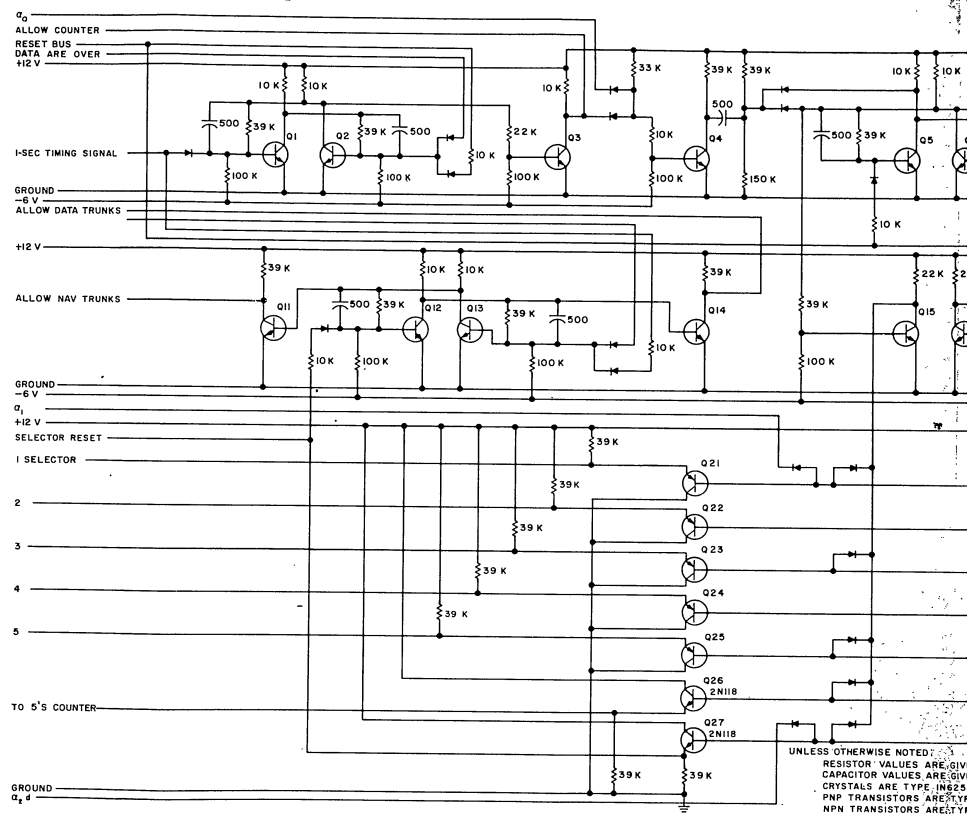
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FIGURE 93
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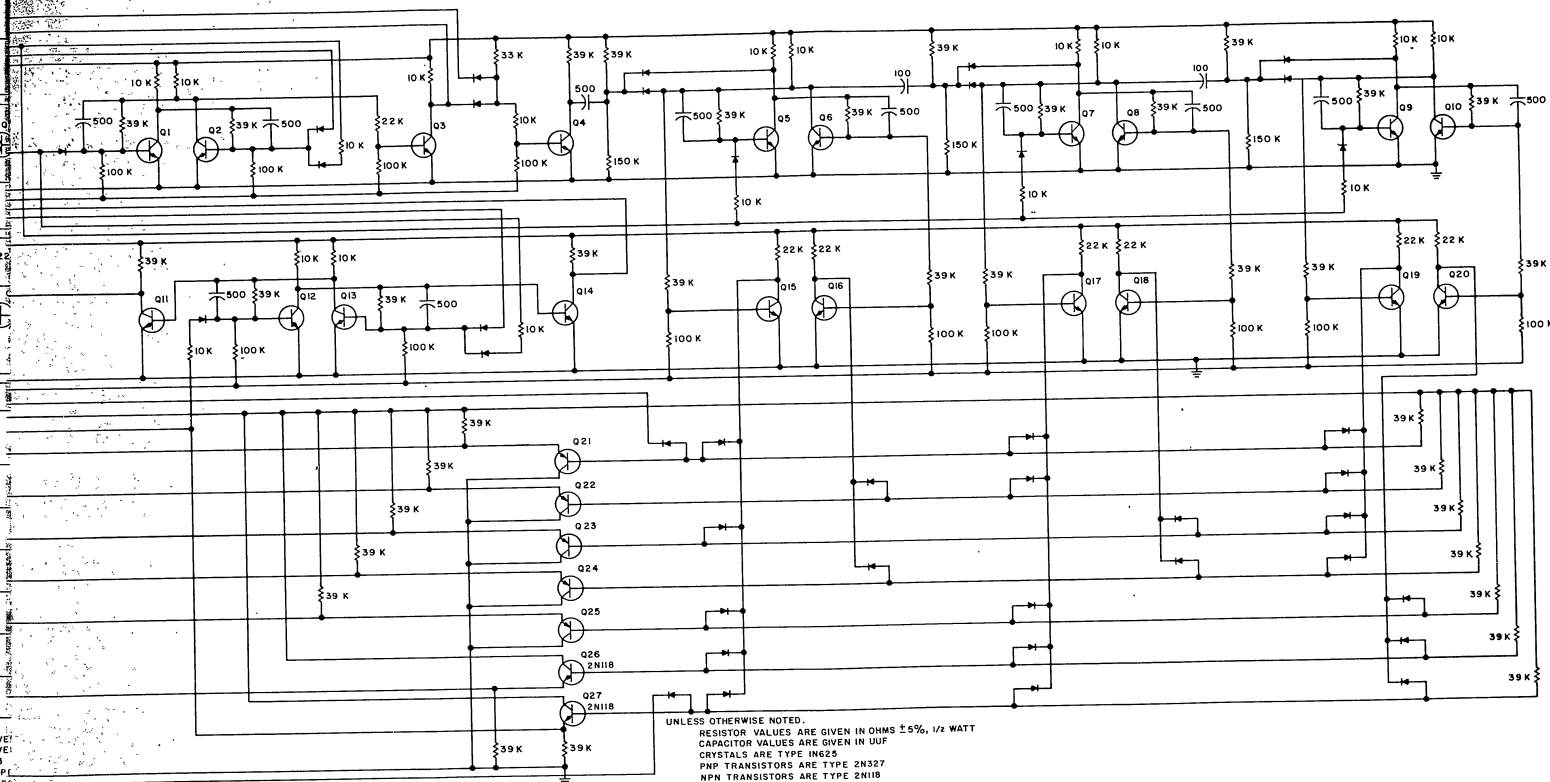


FIGURE 94 SCHEMATIC DIAGRAM OF CONTROL CIRCUITS OF READ-OUT UNIT

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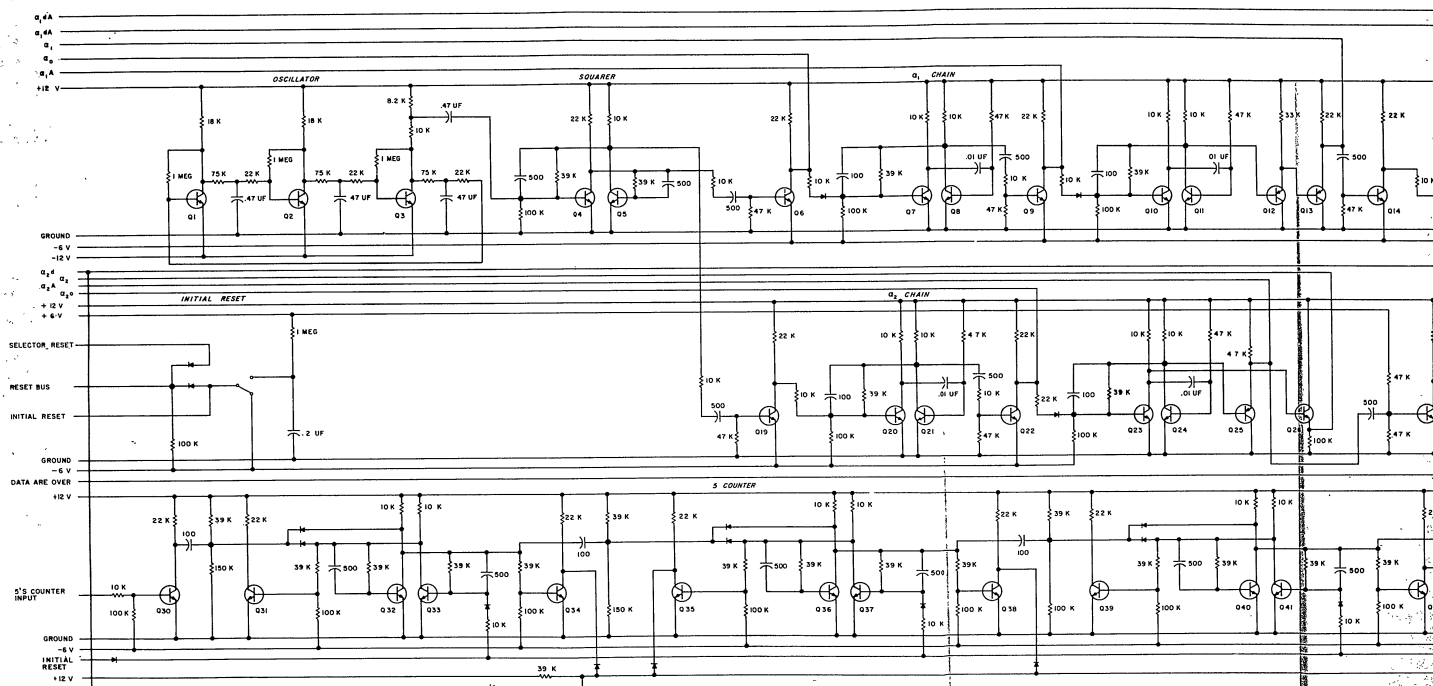
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FIGURE 94
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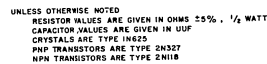


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A. 5'S COUNTER

FIGURE 95. SCHEMATIC DIAGRAM OF 5'S COUNTER AND RECORD GATES OF READ-OUT UNIT

FIGURE 95
SHEET 1 OF 2
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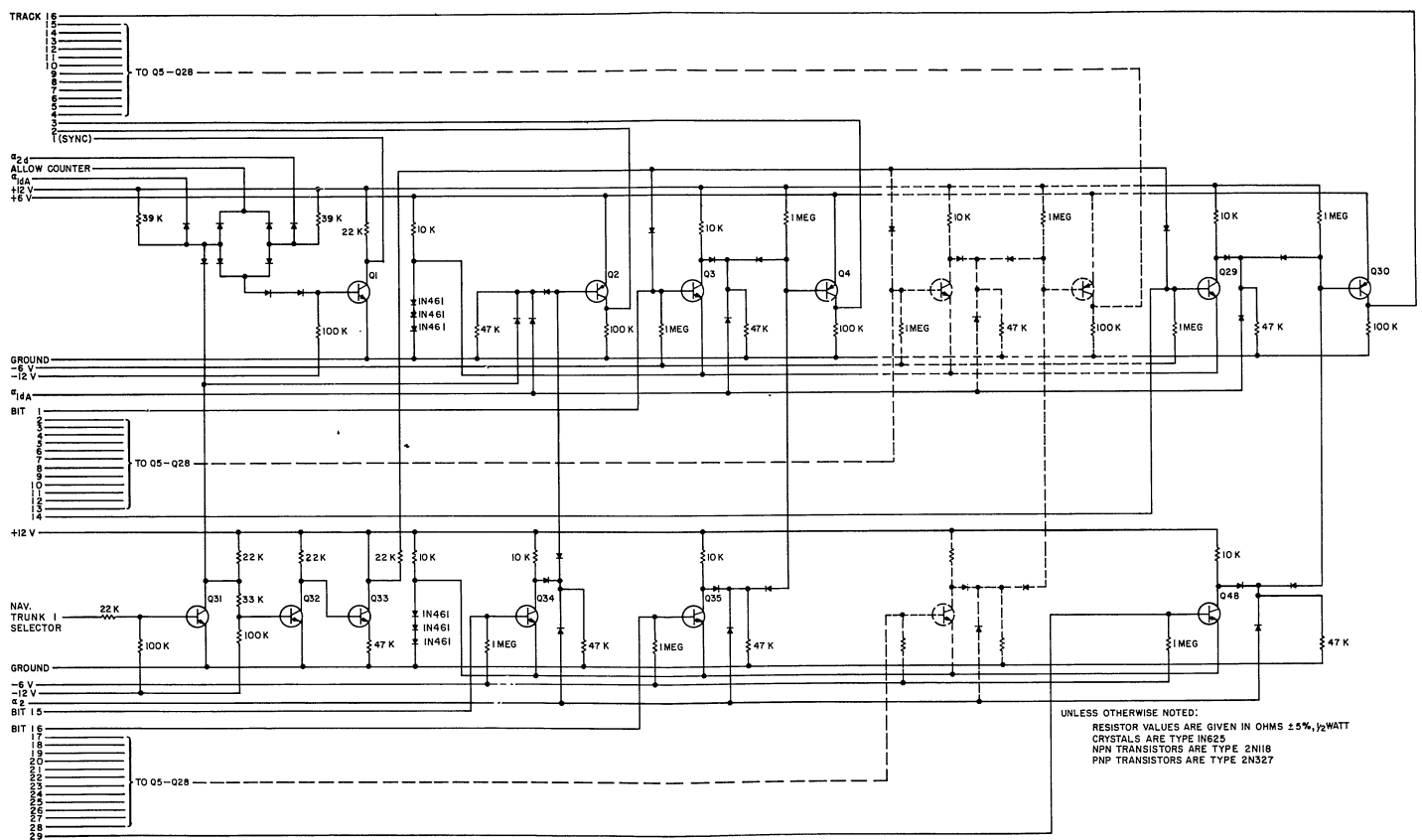
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B. RECORD GATES

FIGURE 95.

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SHEET 2 OF 2
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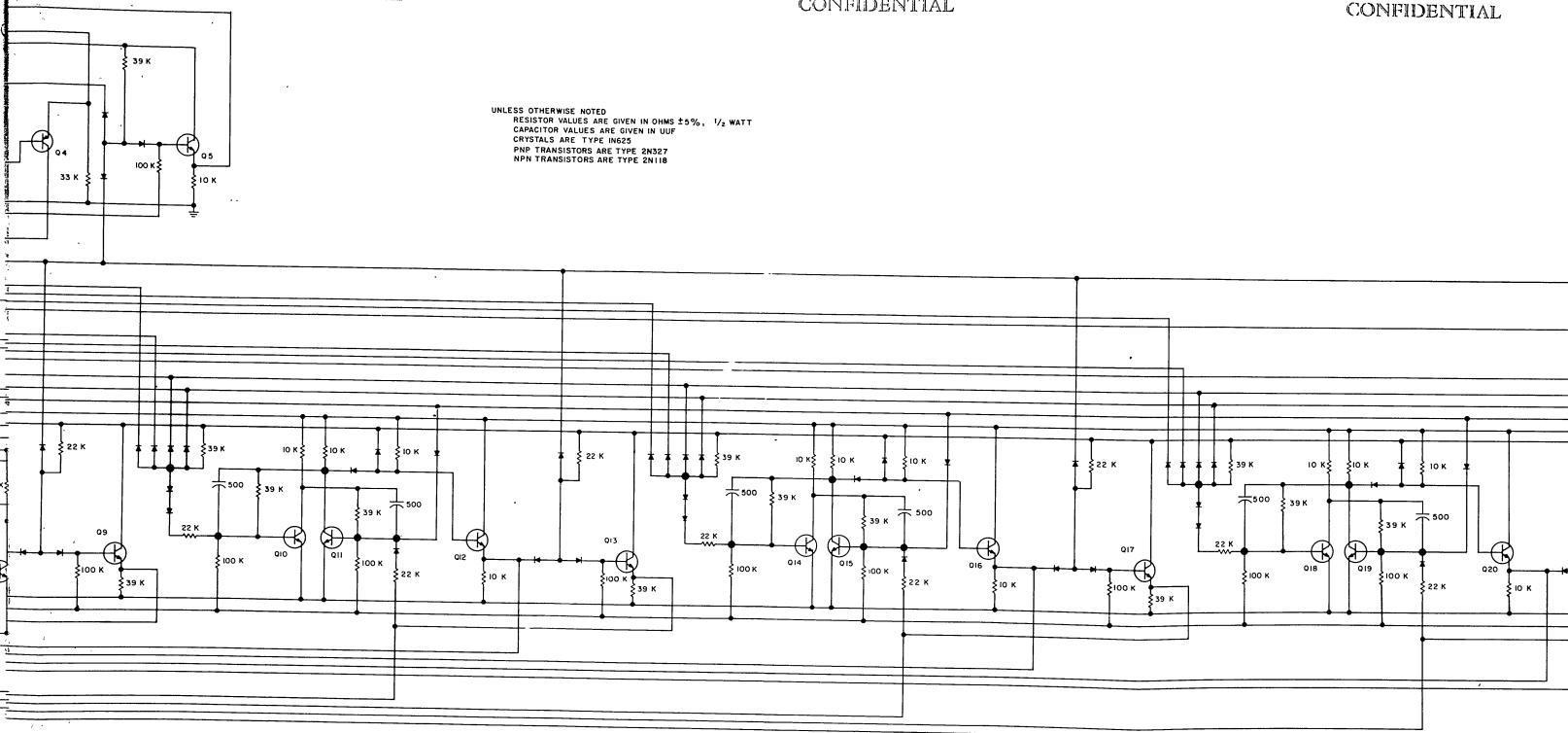


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UNLESS OTHERWISE NOTED
RESISTOR VALUES ARE GIVEN IN OHMS $\pm 5\%$, $\frac{1}{2}$ WATT
CAPACITOR VALUES ARE GIVEN IN UUF
CRYSTALS ARE TYPE IN25
PNP TRANSISTORS ARE TYPE 2N327
NPN TRANSISTORS ARE TYPE 2N118



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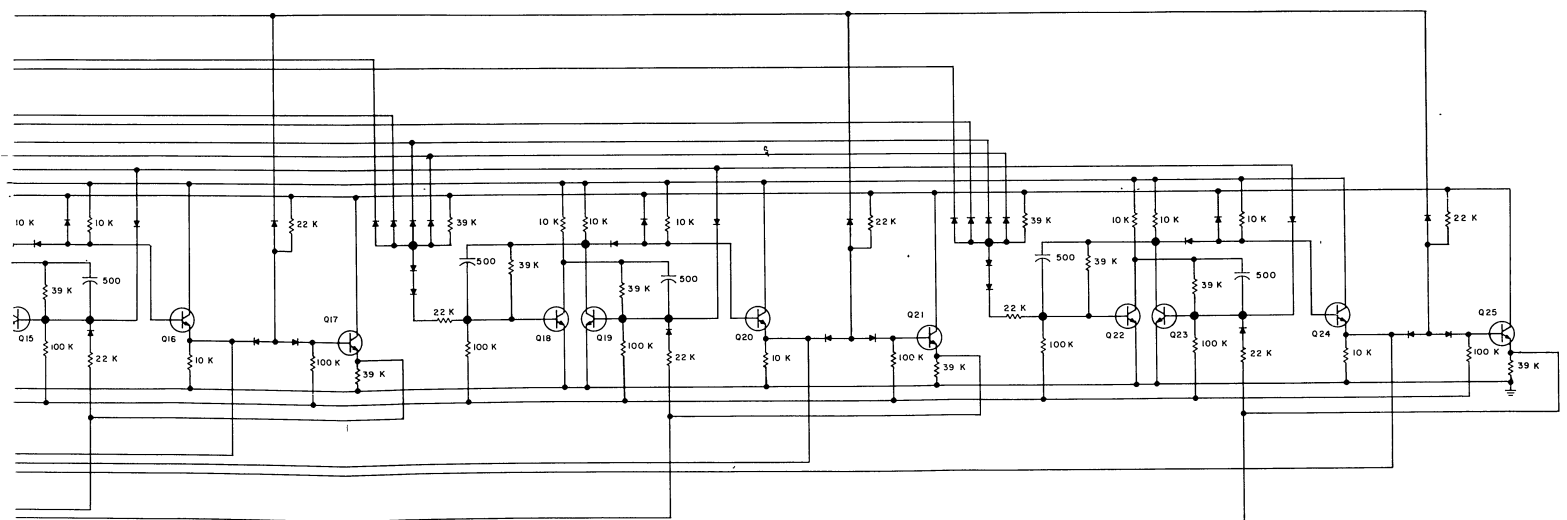


FIGURE 96. SCHEMATIC DIAGRAM OF TRUNK GATES FOR READ-OUT UNIT

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FIGURE 96
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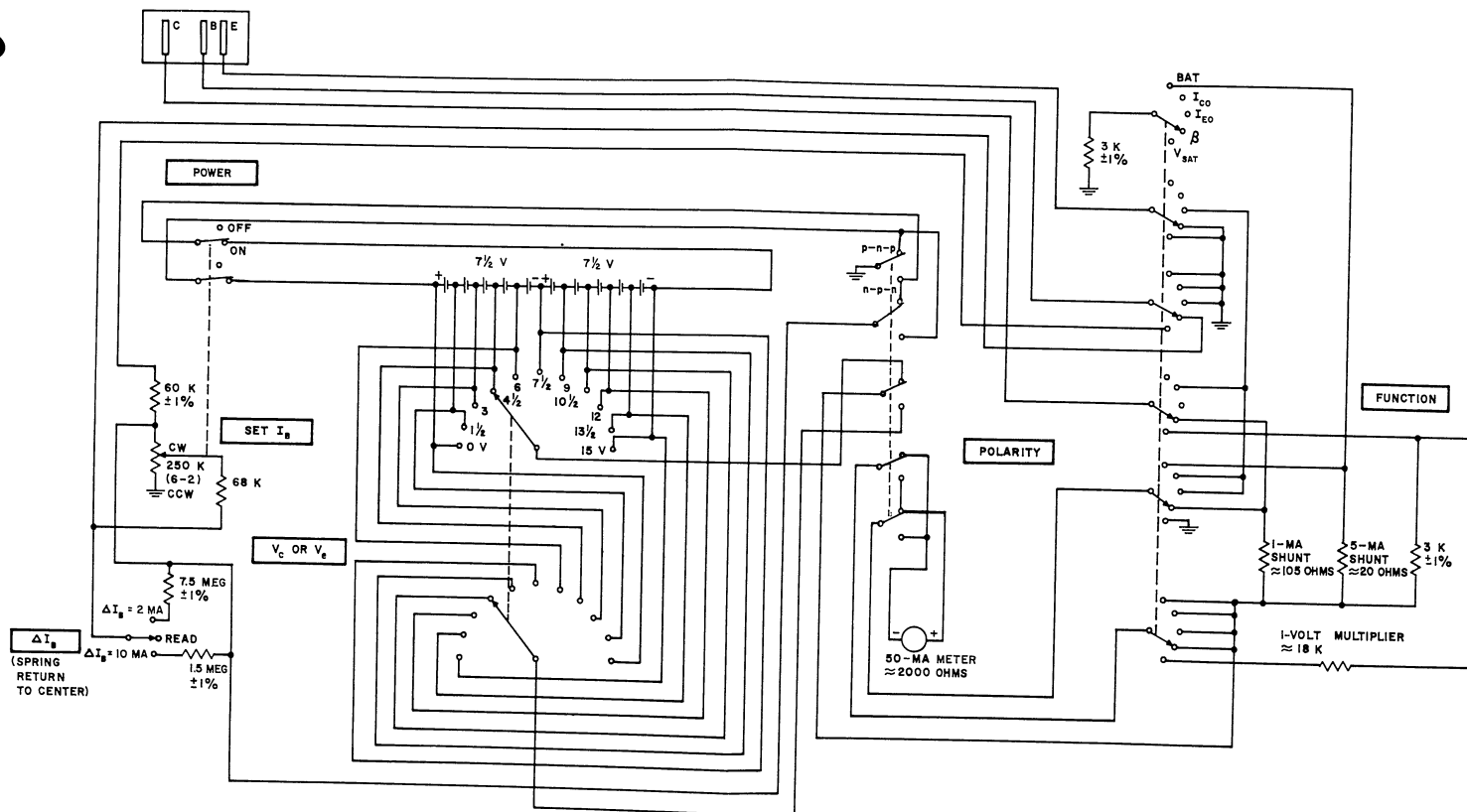


FIGURE 97. SCHEMATIC DIAGRAM OF JUNCTION-TRANSISTOR TESTER

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FIGURE 97
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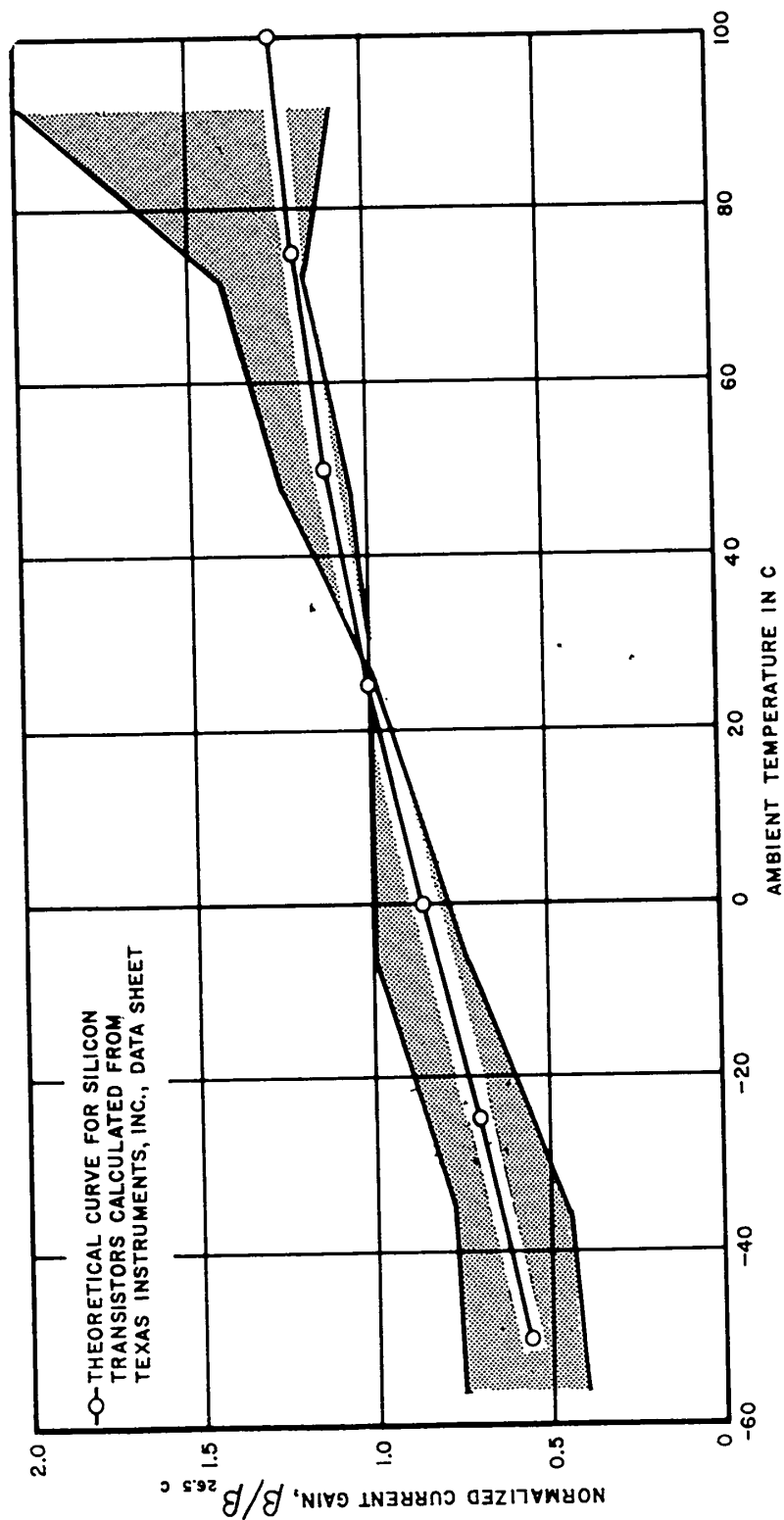


FIGURE 98. SPREAD OF NORMALIZED COLLECTOR-TO-BASE CURRENT GAIN VS AMBIENT TEMPERATURE OF TRANSISTORS

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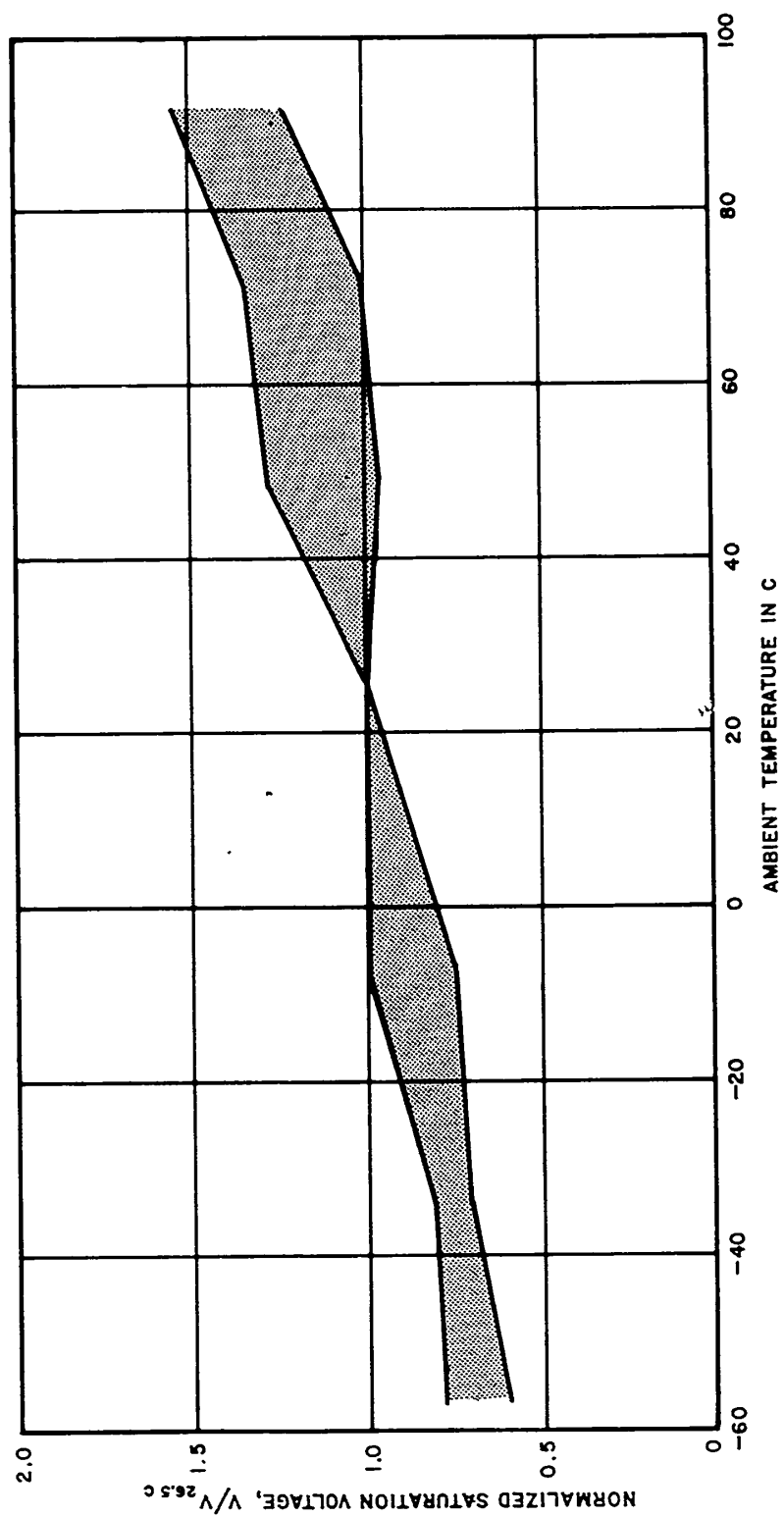
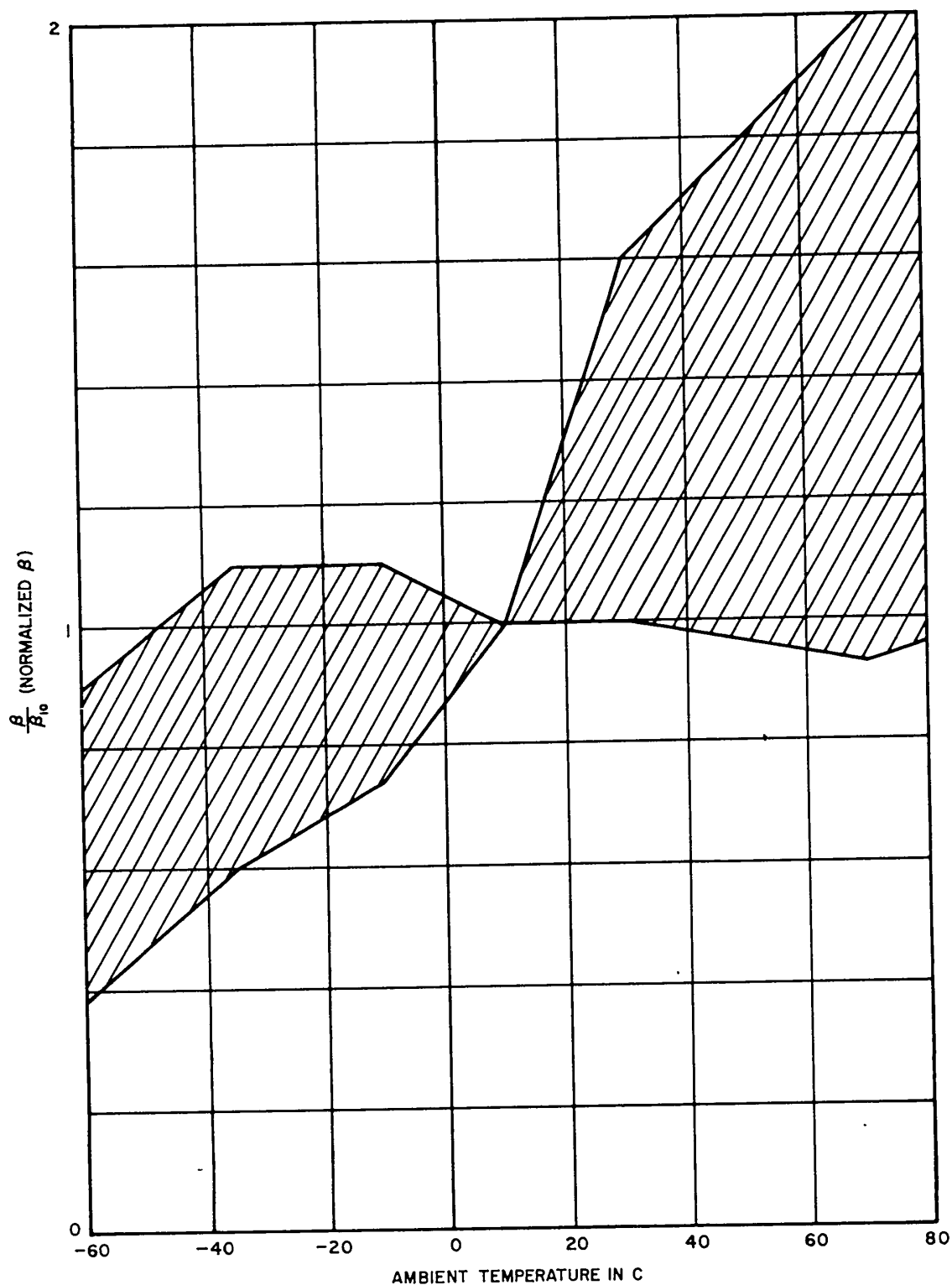


FIGURE 99. SPREAD OF NORMALIZED SATURATION VOLTAGE VS AMBIENT TEMPERATURE OF TRANSISTORS

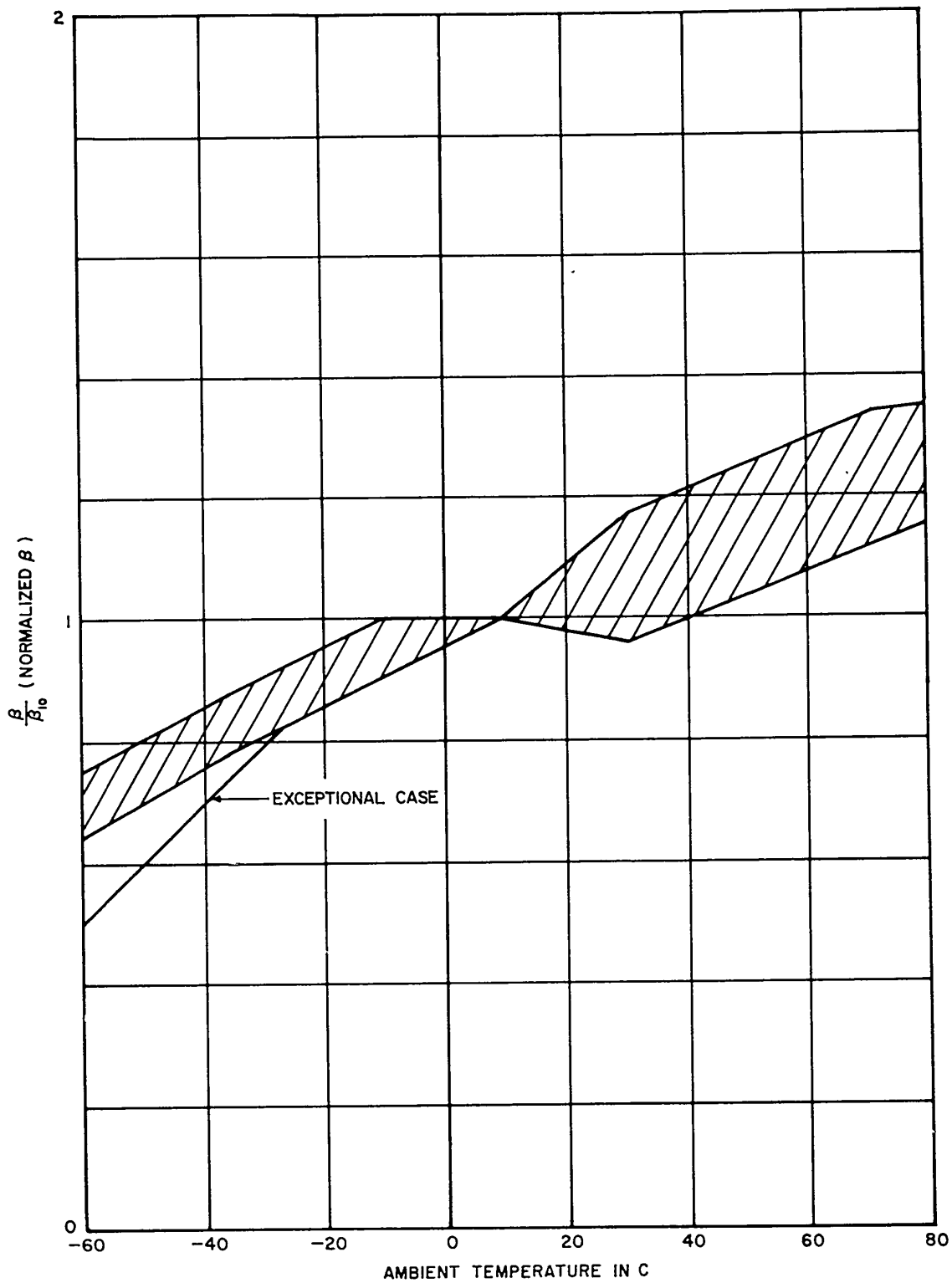
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FIGURE 100. NORMALIZED β VS TEMPERATURE OF TYPE 903 TRANSISTOR

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FIGURE 101. NORMALIZED β VS TEMPERATURE OF TYPE 904 TRANSISTOR

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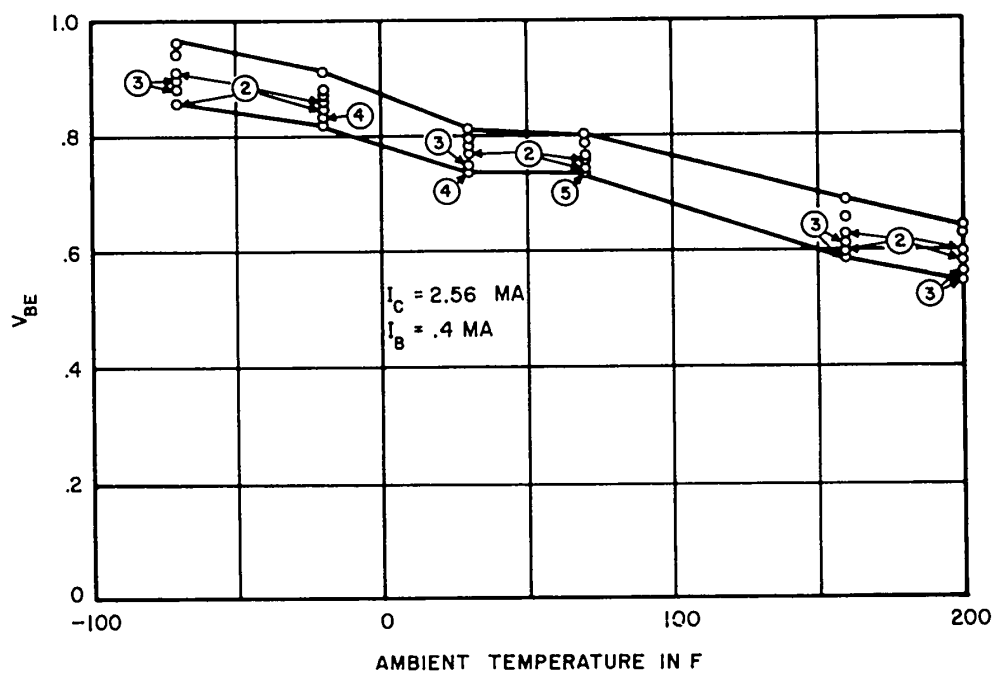


FIGURE 102. BASE SATURATION VOLTAGE VS TEMPERATURE FOR TWELVE 2N118 TRANSISTORS

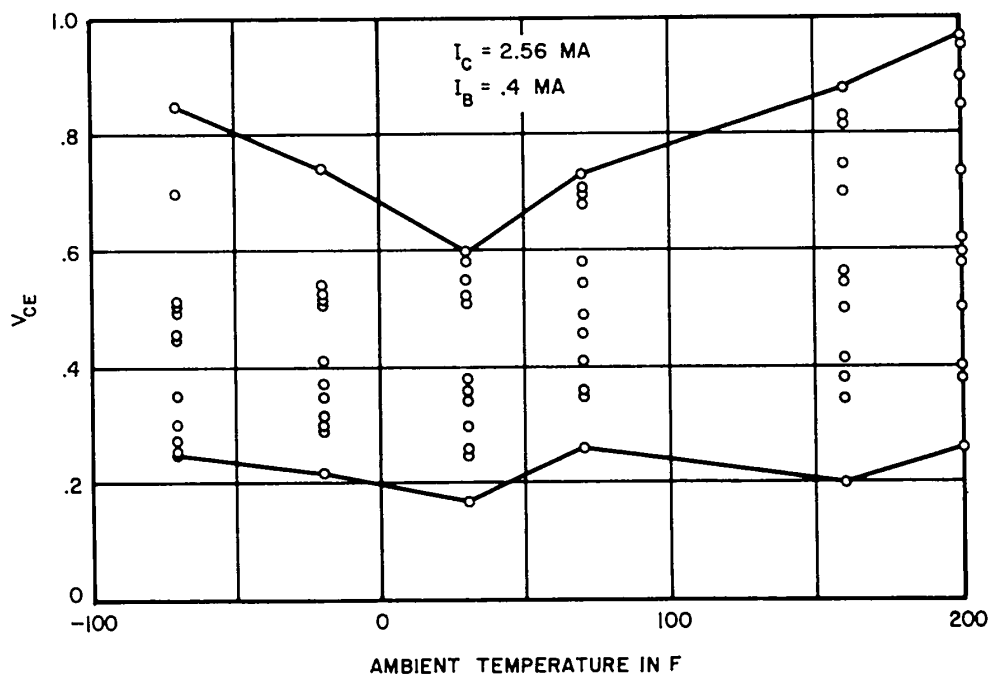


FIGURE 103. COLLECTOR SATURATION VOLTAGE VS TEMPERATURE FOR TWELVE 2N118 TRANSISTORS

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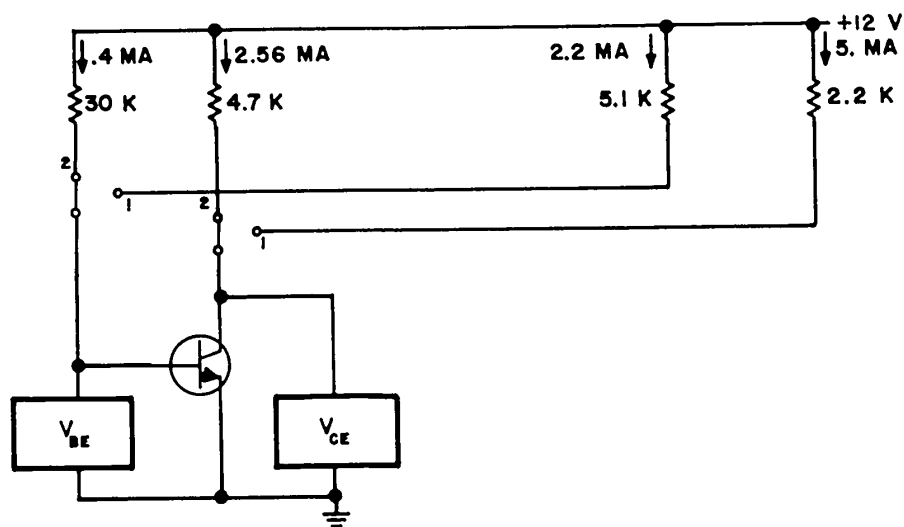


FIGURE 104. SCHEMATIC DIAGRAM OF TEST CIRCUIT FOR SATURATION-VOLTAGE MEASUREMENTS

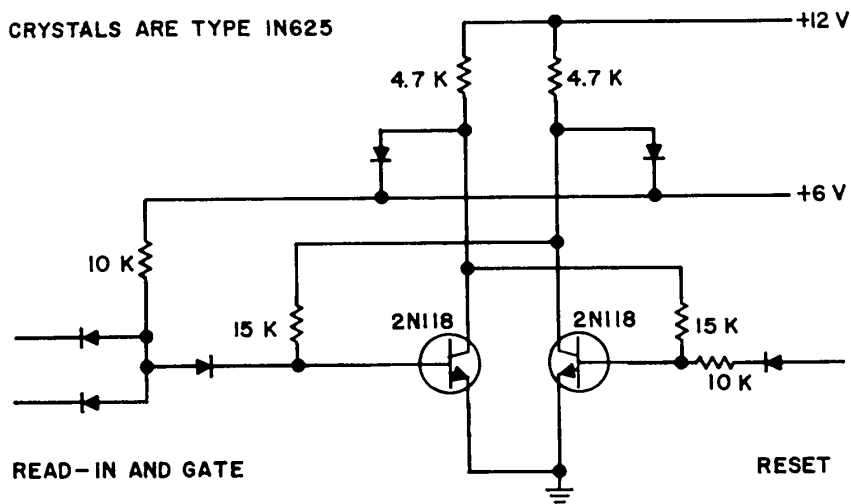


FIGURE 105. BASIC FLIP-FLOP

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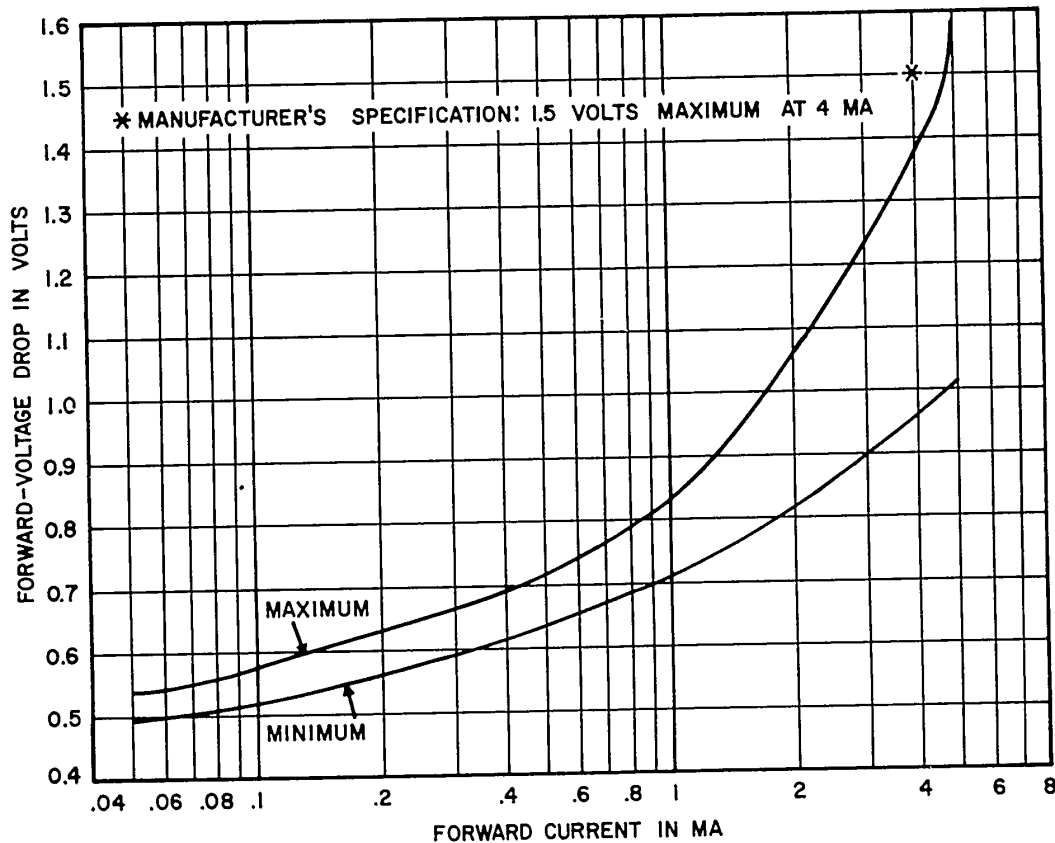


FIGURE 106. IN625 DIODE FORWARD VOLTAGE VS CURRENT

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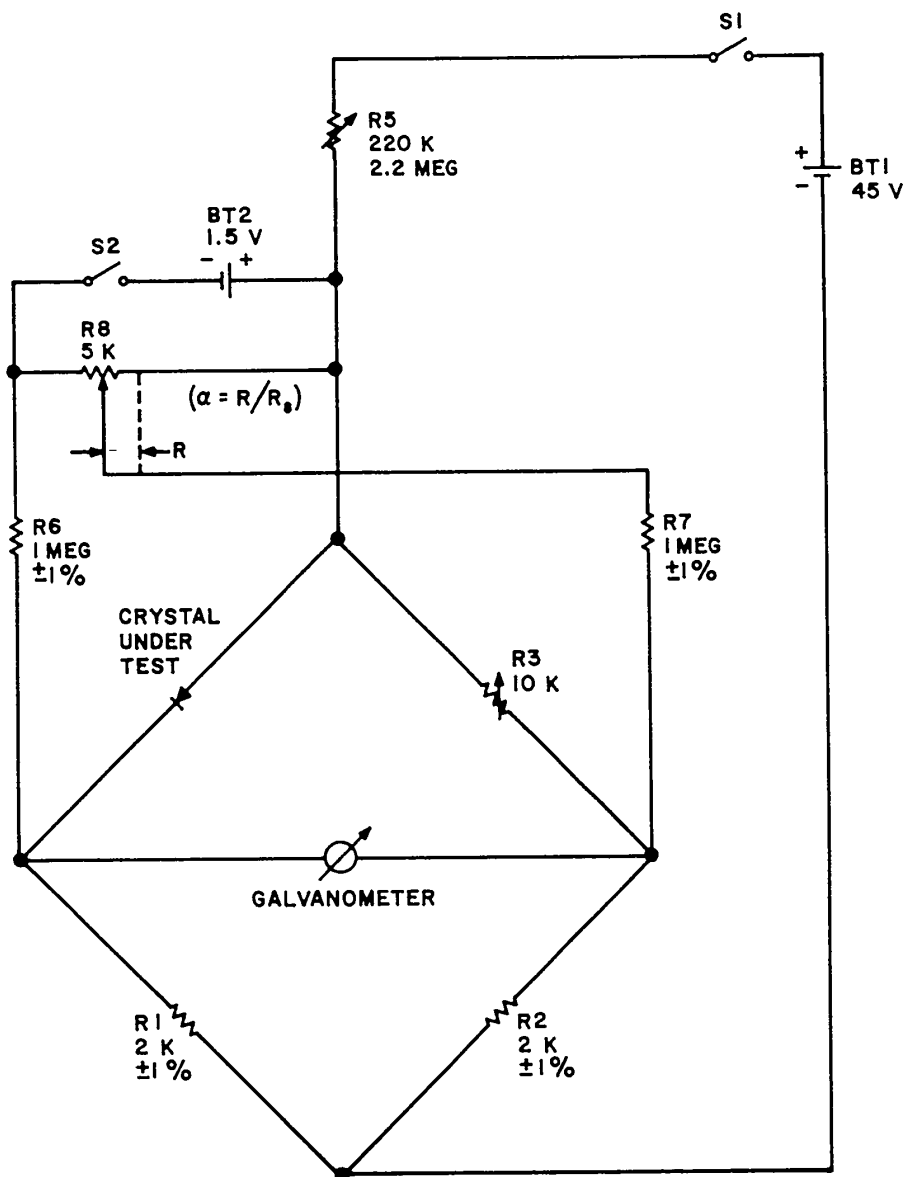


FIGURE 107. SCHEMATIC DIAGRAM OF TESTER FOR VIDEO-DETECTOR CRYSTALS

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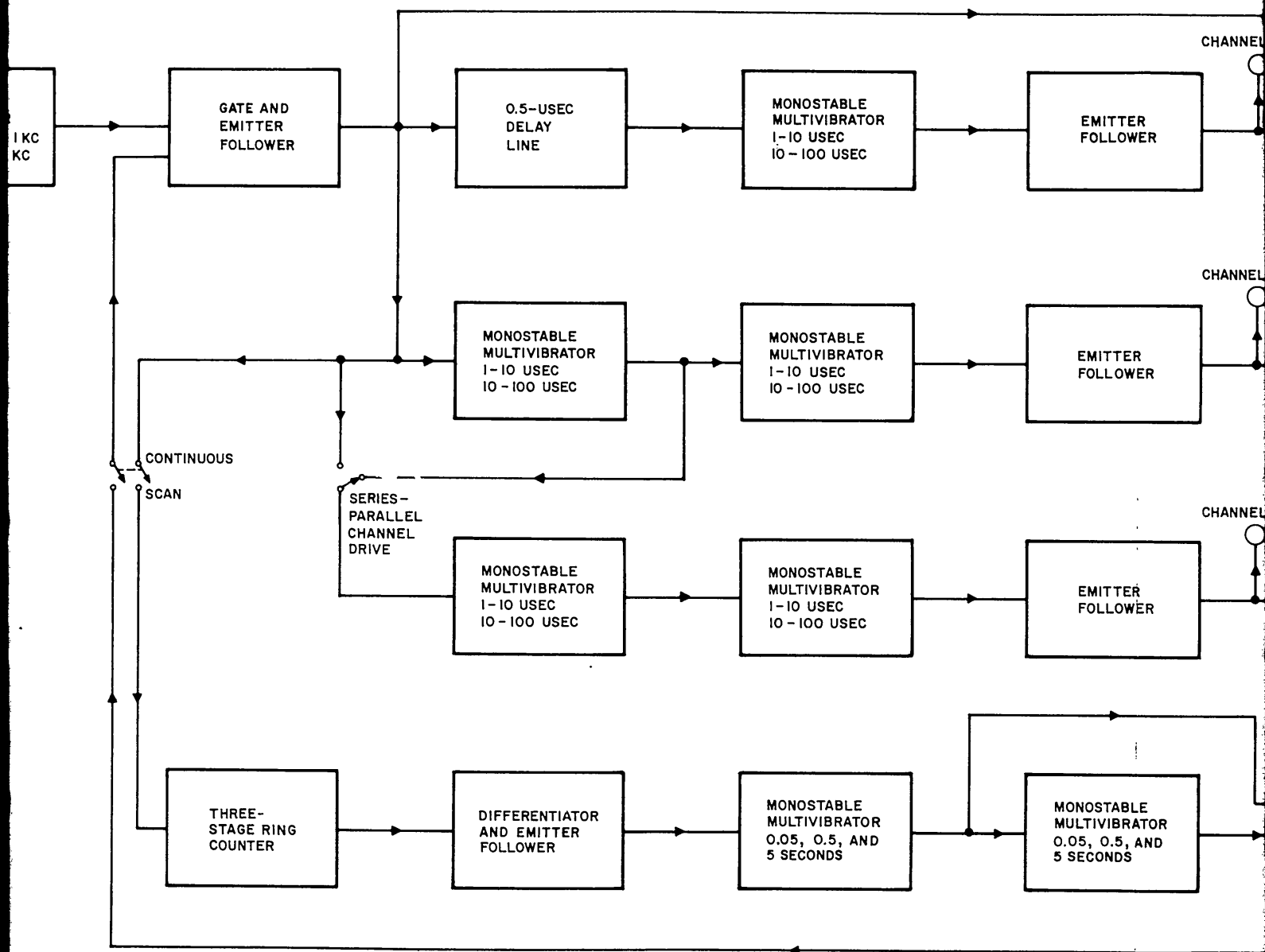


FIGURE 108. BLOCK DIA

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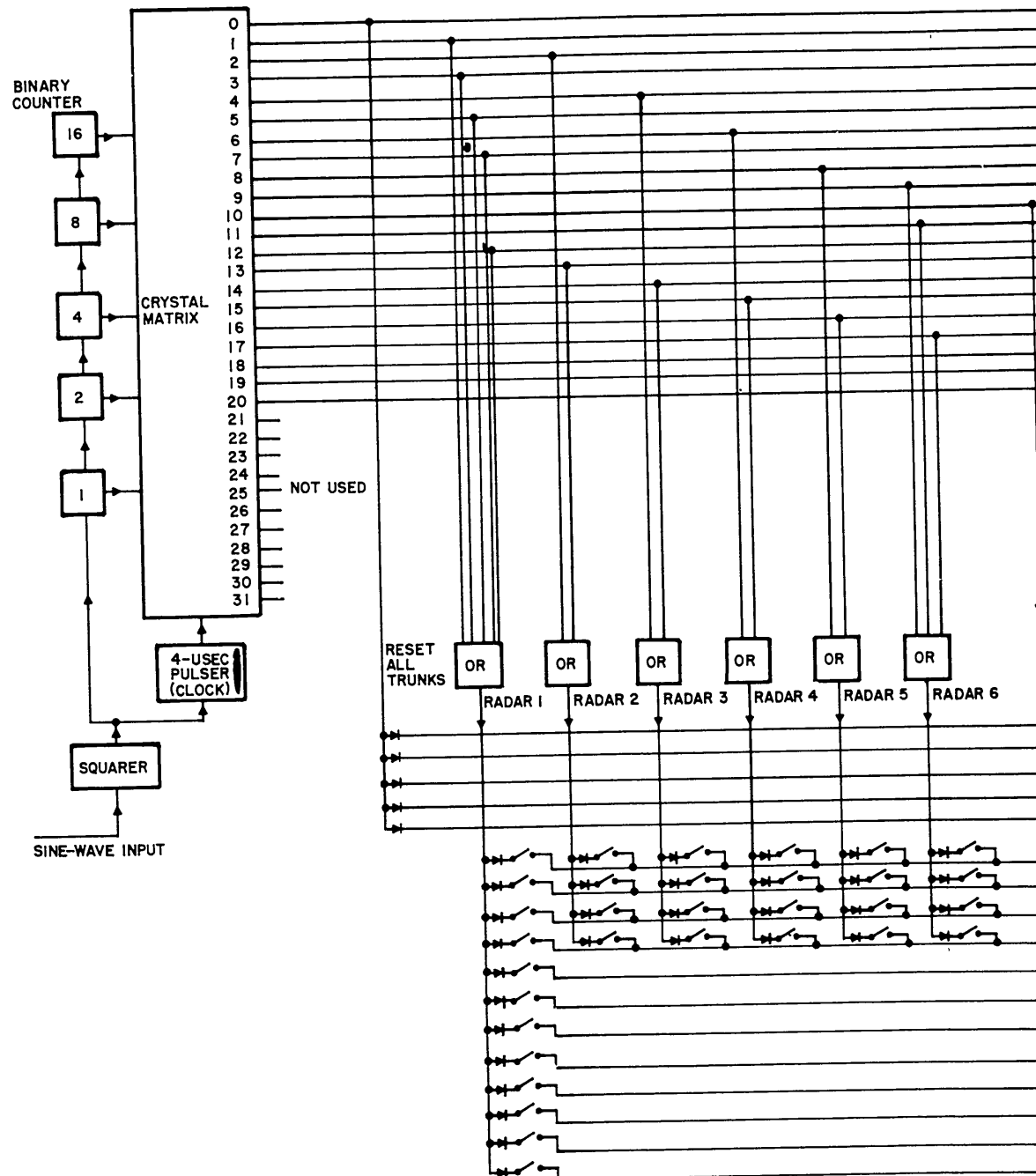


FIGURE 109. BLO

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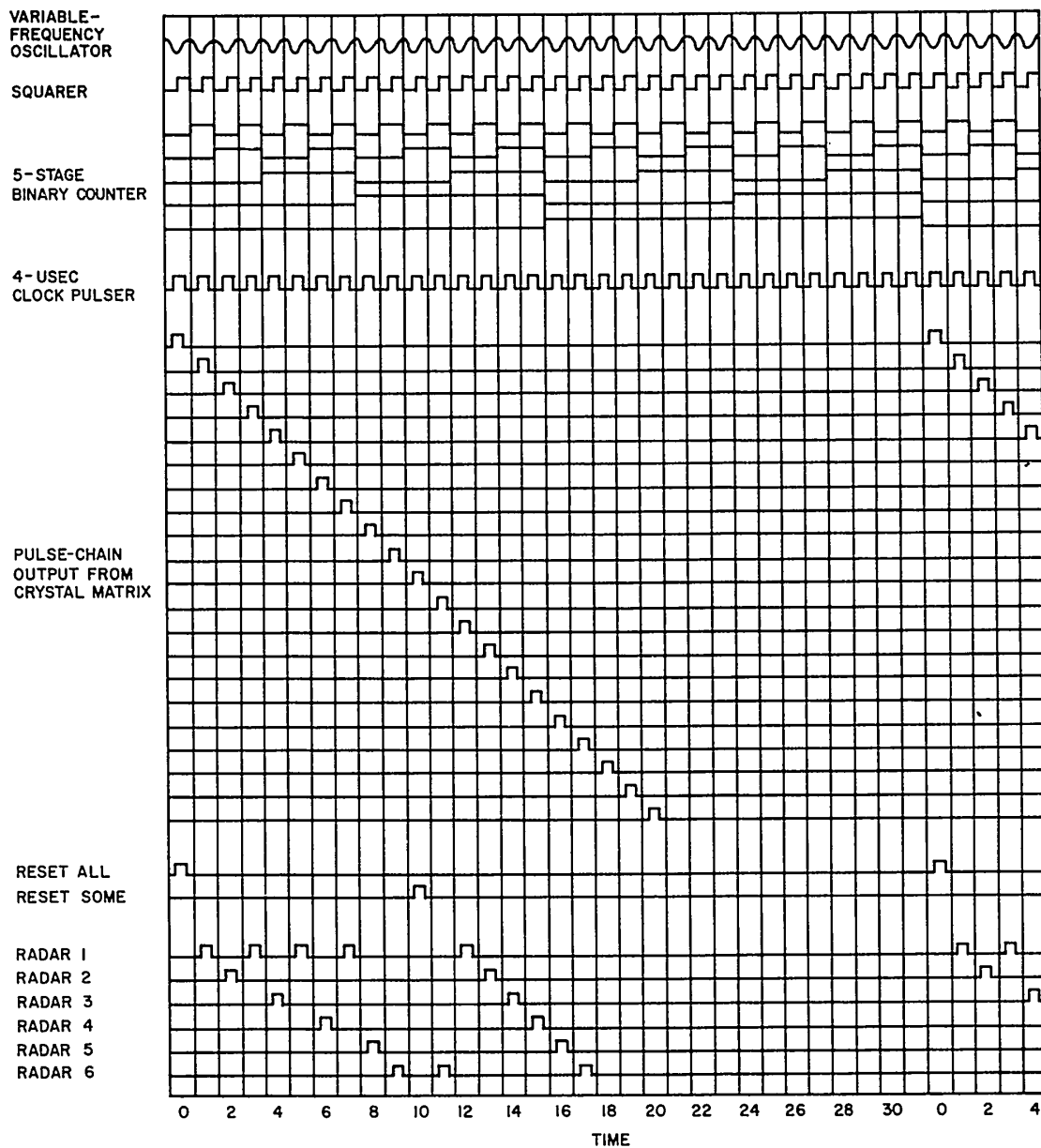
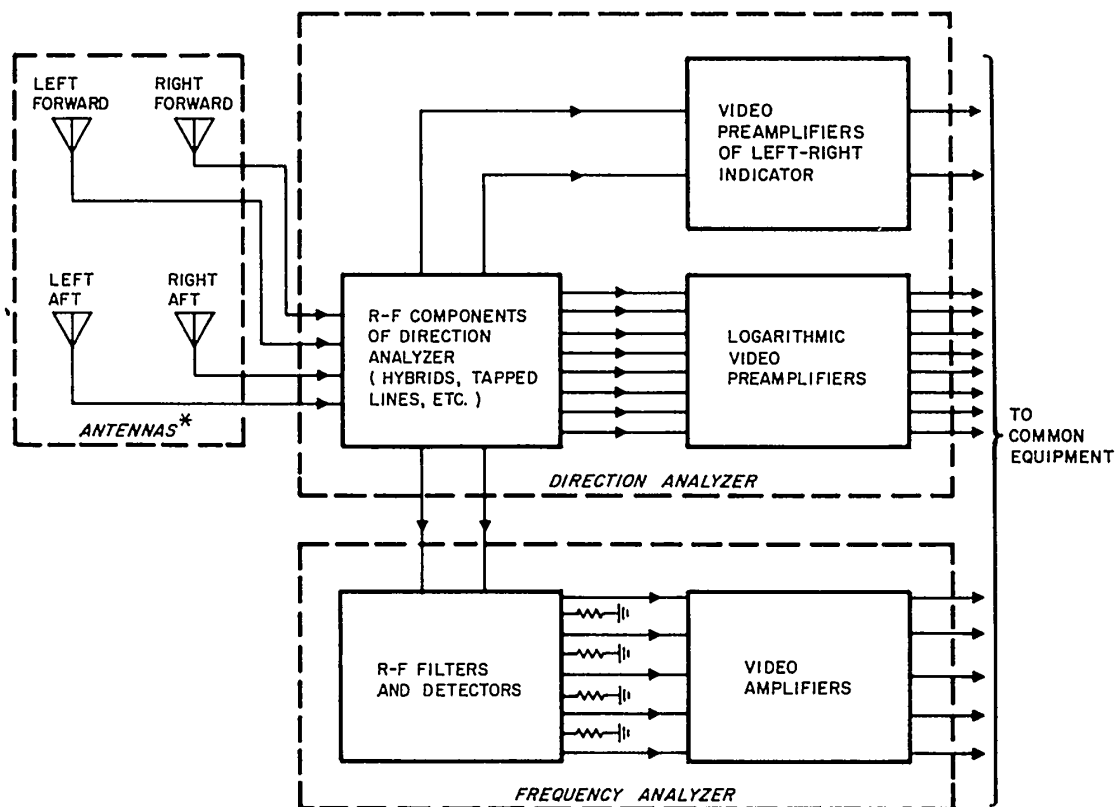


FIGURE IIO. TIMING DIAGRAM OF TEST-PROBLEM GENERATOR

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* ANTENNAS WERE 1/8-SCALE MODELS OF ANTENNAS FOR SUB-BAND 2, MOUNTED ON A 1/8-SCALE MODEL B-47, AND FULL-SCALE ANTENNAS FOR SUB-BAND 5 MOUNTED ON A MOCK-UP OF A SECTION OF A B-47.

FIGURE III. BLOCK DIAGRAM OF SUB-BAND 5 PORTION OF FEASIBILITY-DEMONSTRATION SYSTEM

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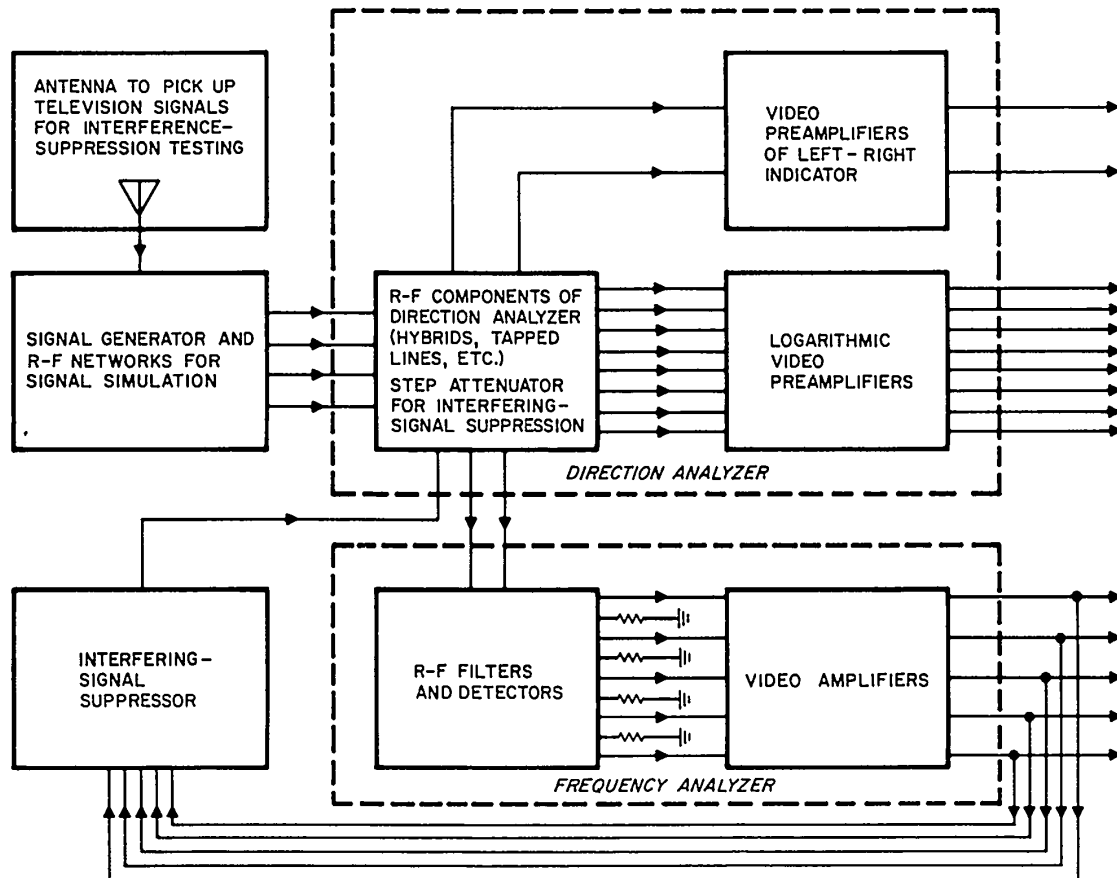


FIGURE 112. BLOCK DIAGRAM OF SUB-BAND 2 PORTION OF
FEASIBILITY-DEMONSTRATION SYSTEM

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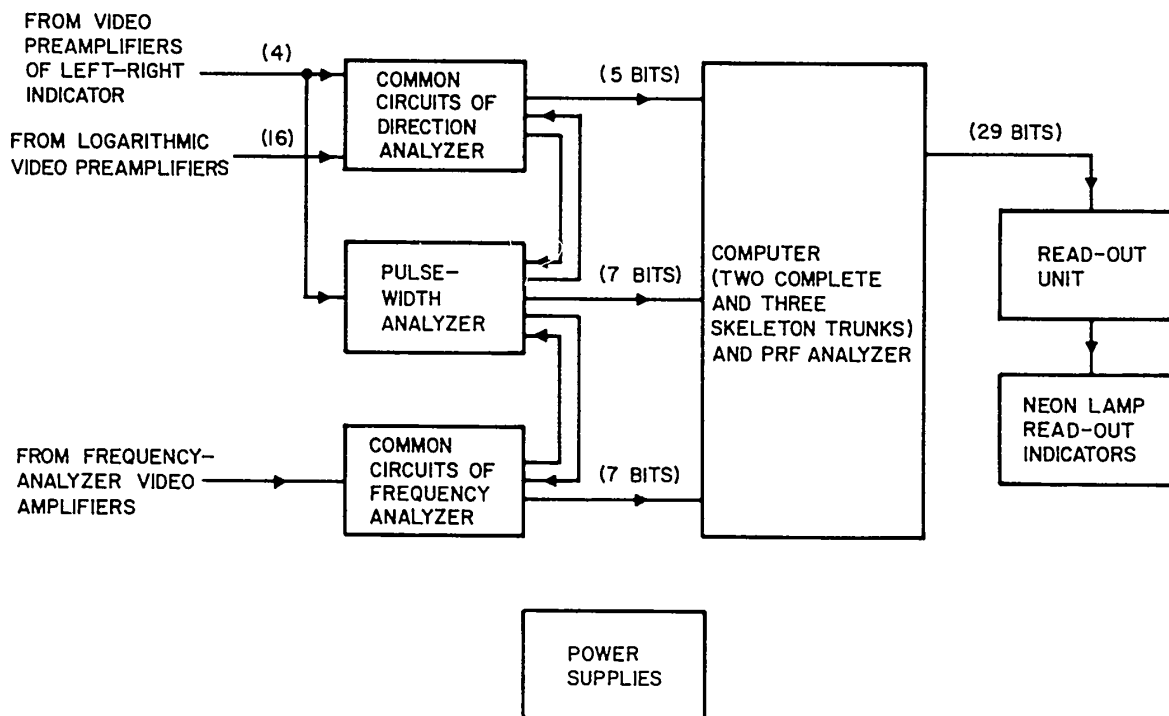


FIGURE 113. BLOCK DIAGRAM OF COMMON EQUIPMENT FOR FEASIBILITY-DEMONSTRATION SYSTEM

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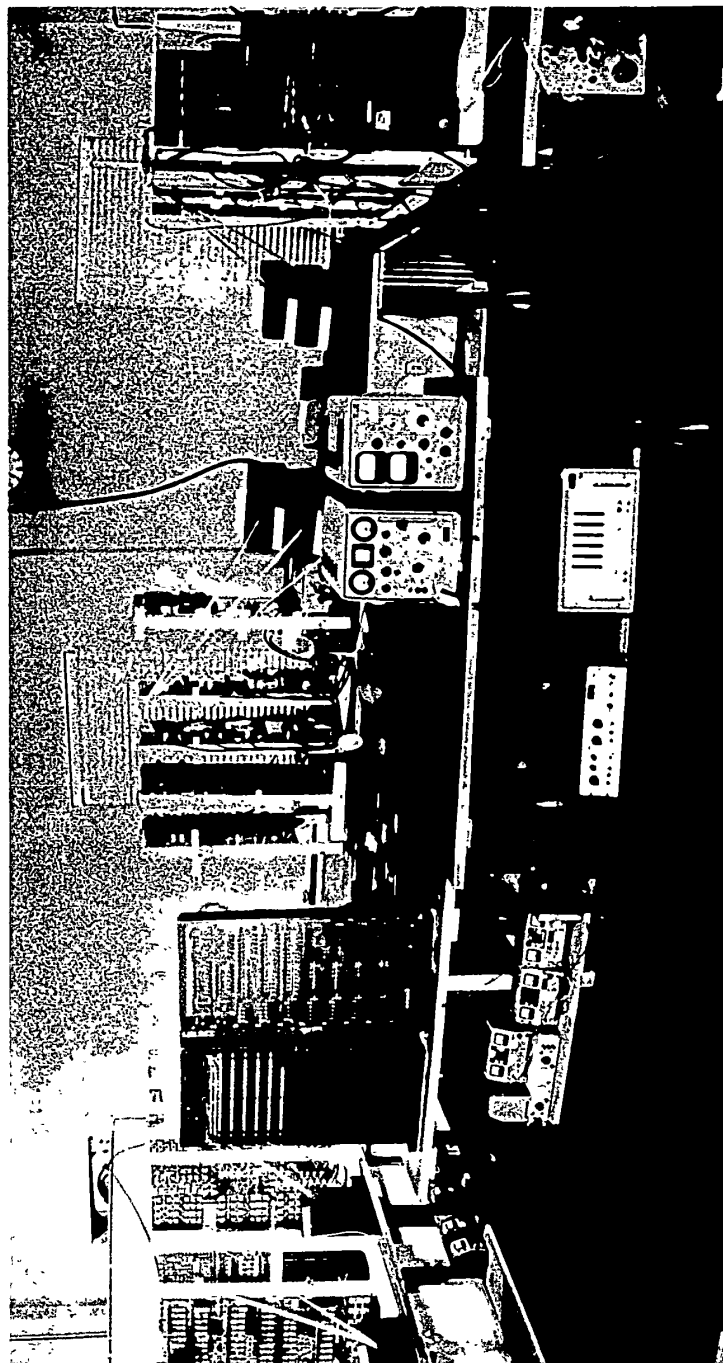


FIGURE 114. GENERAL VIEW OF FEASIBILITY-DEMONSTRATION SYSTEM

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AN/DLD-2
OUTPUT INDICATOR

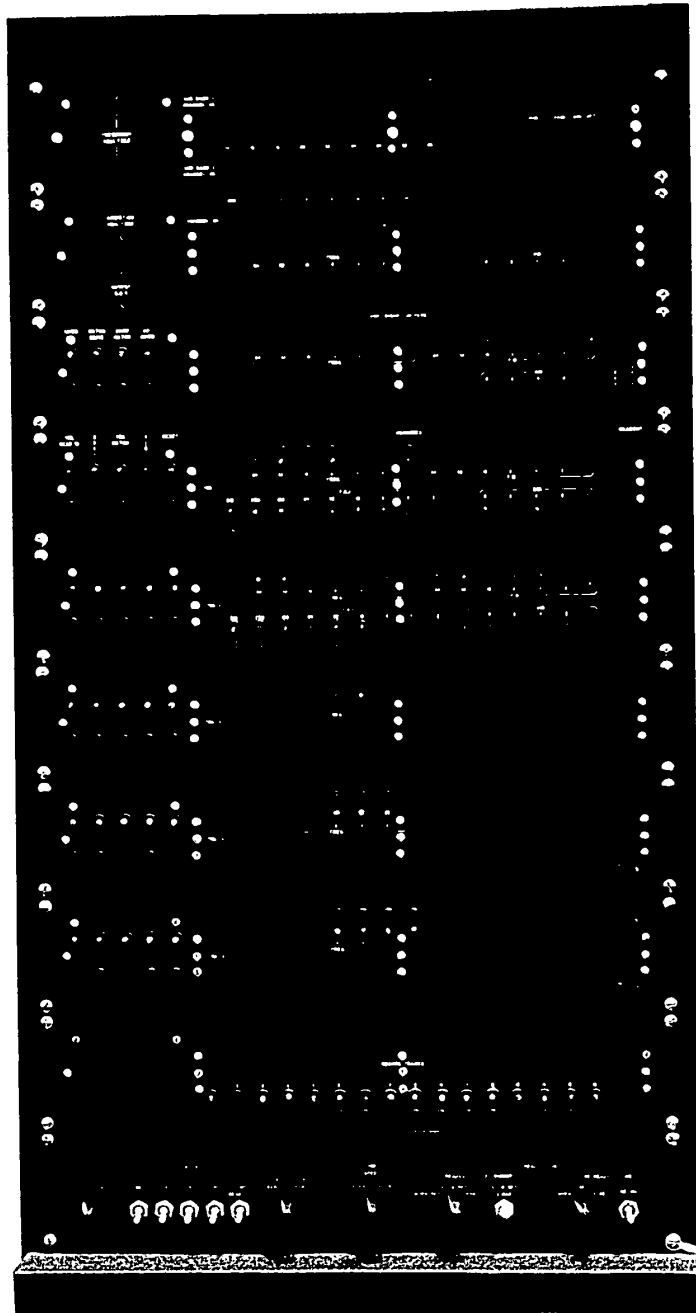


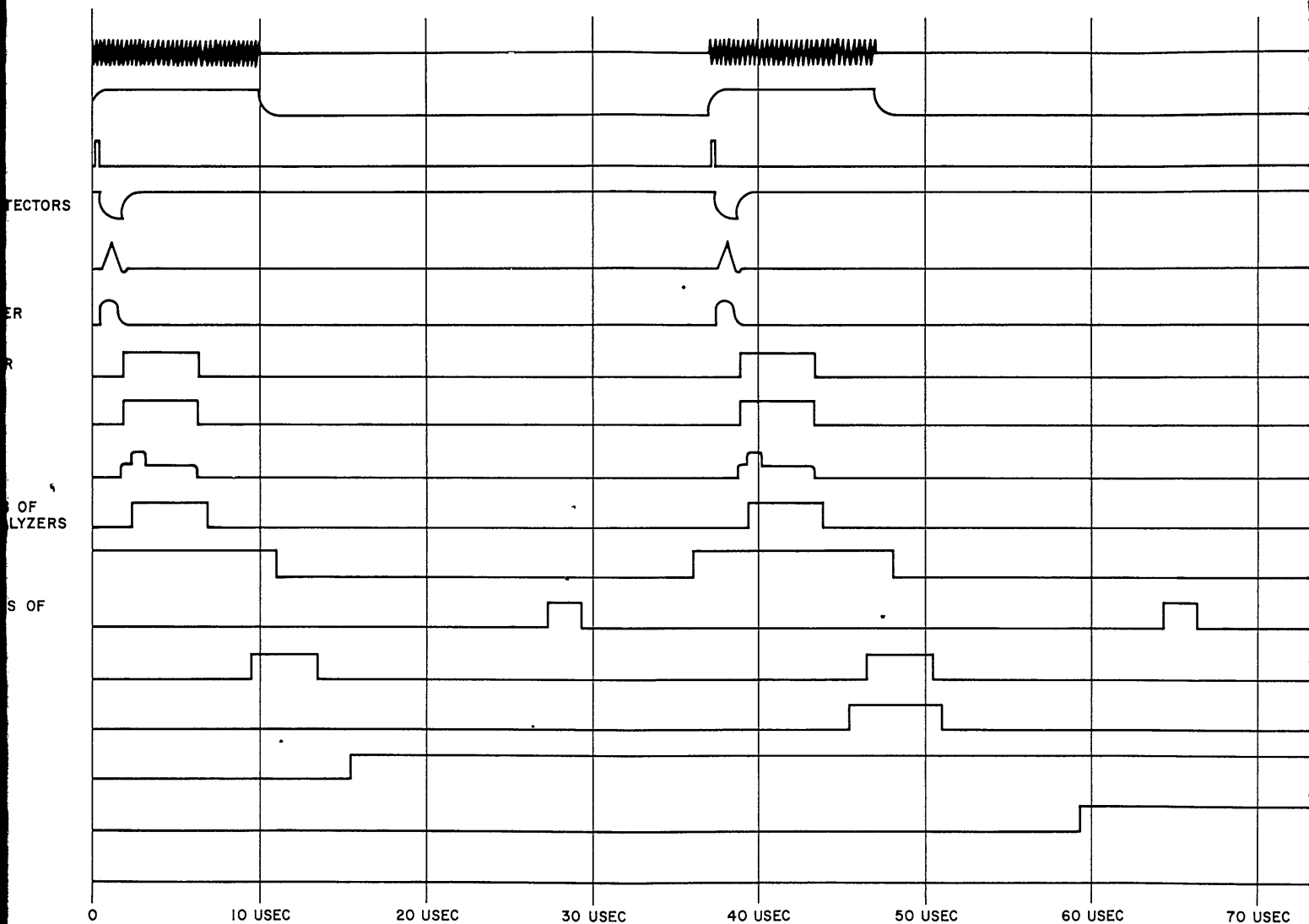
FIGURE 115. NEON-LAMP DISPLAY FOR READ-OUT OF FEASIBILITY-DEMONSTRATION SYSTEM

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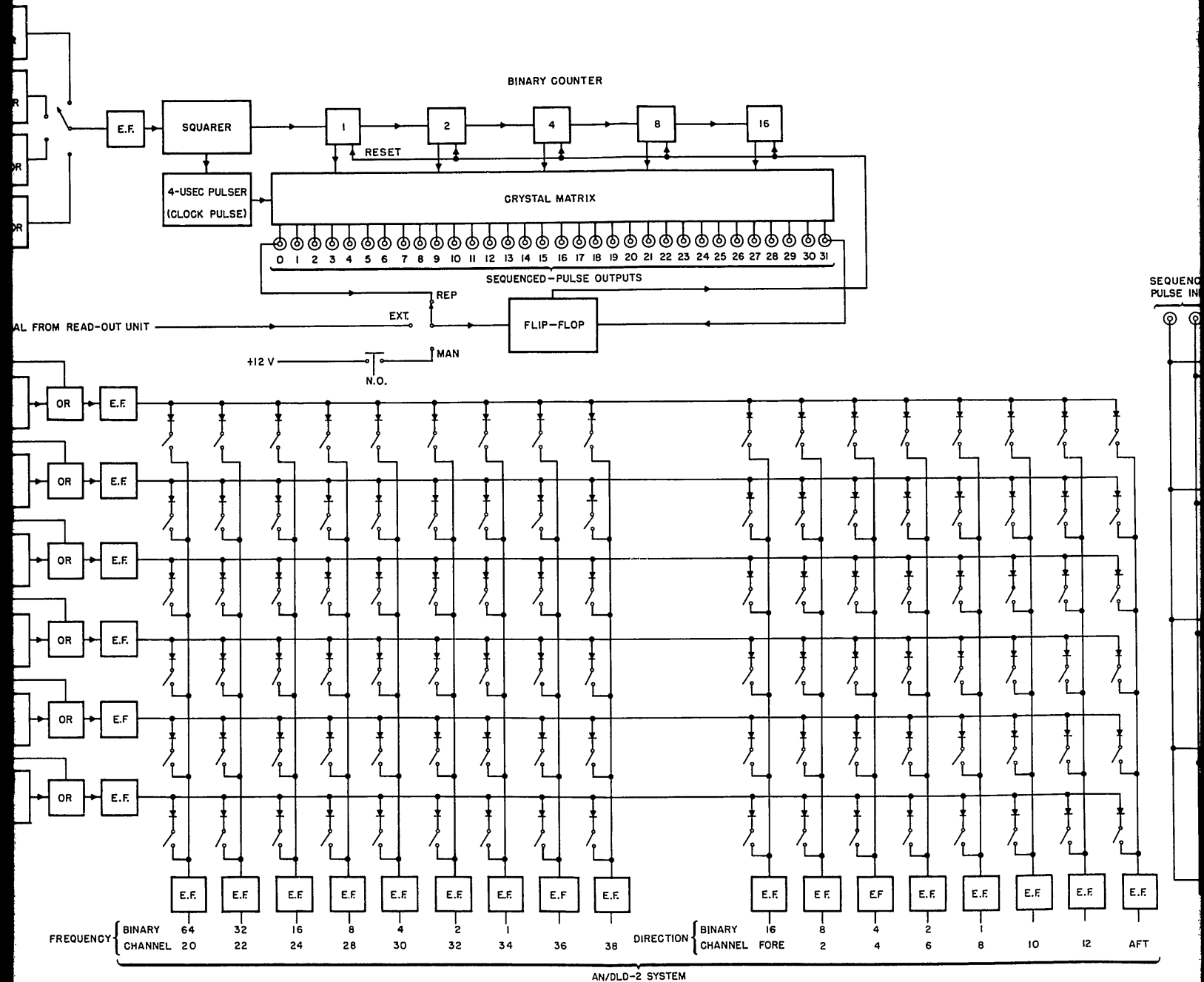
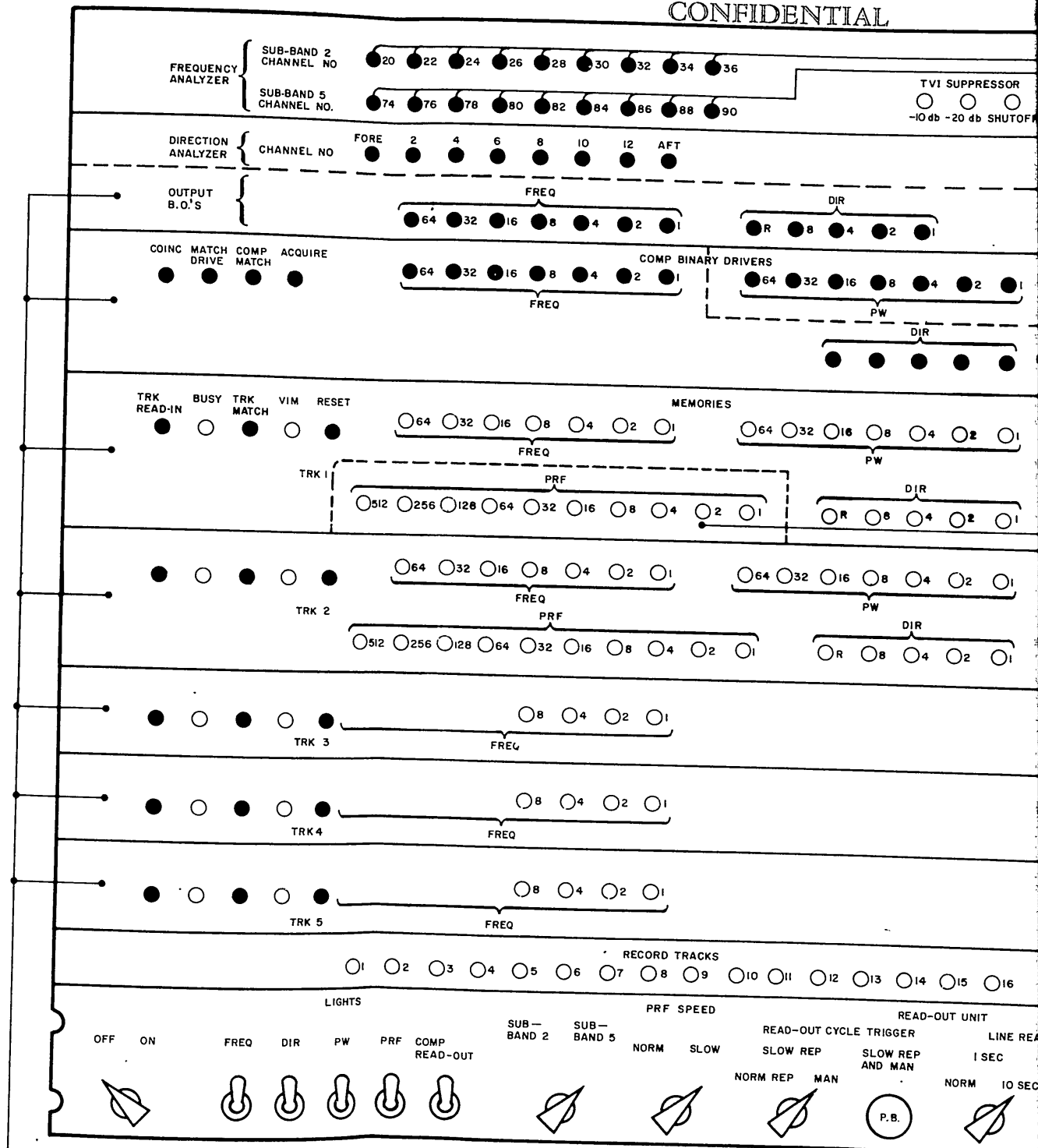


FIGURE 117. BLOCK DIAGRAM OF TEST-PROBLEM GENERATOR FOR FE

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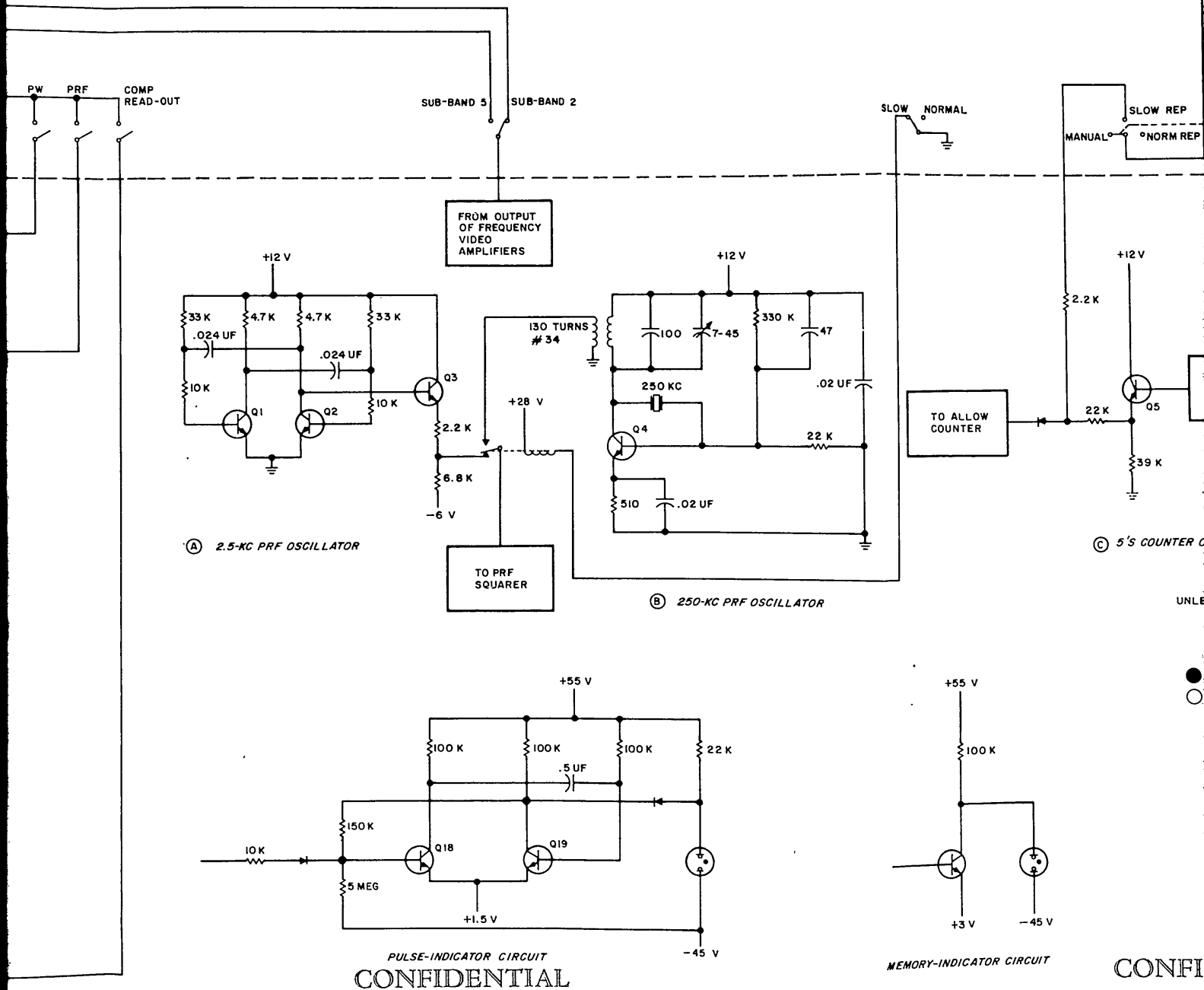


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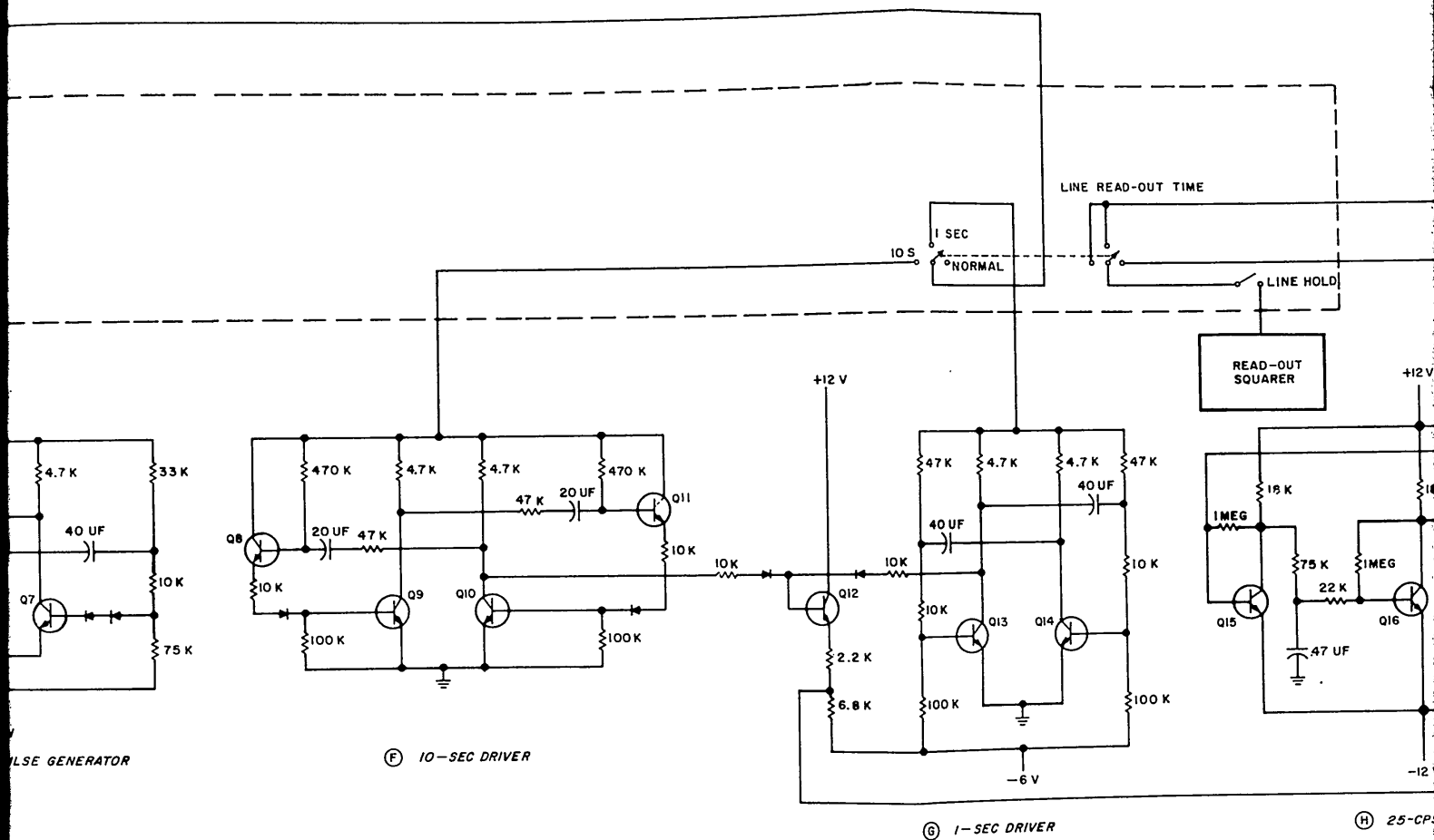


FIGURE 118. PARTIAL SCHEMATIC

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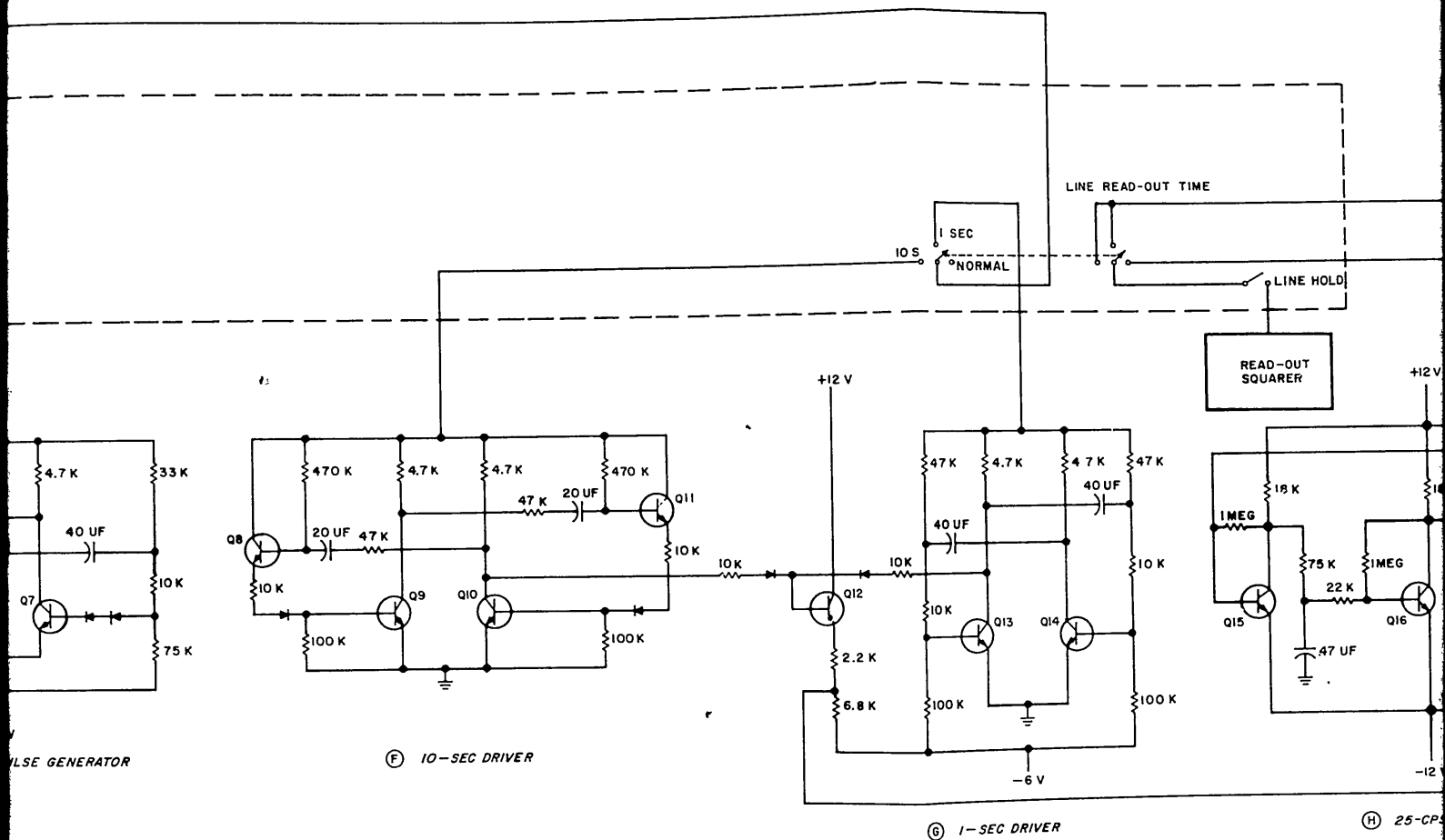


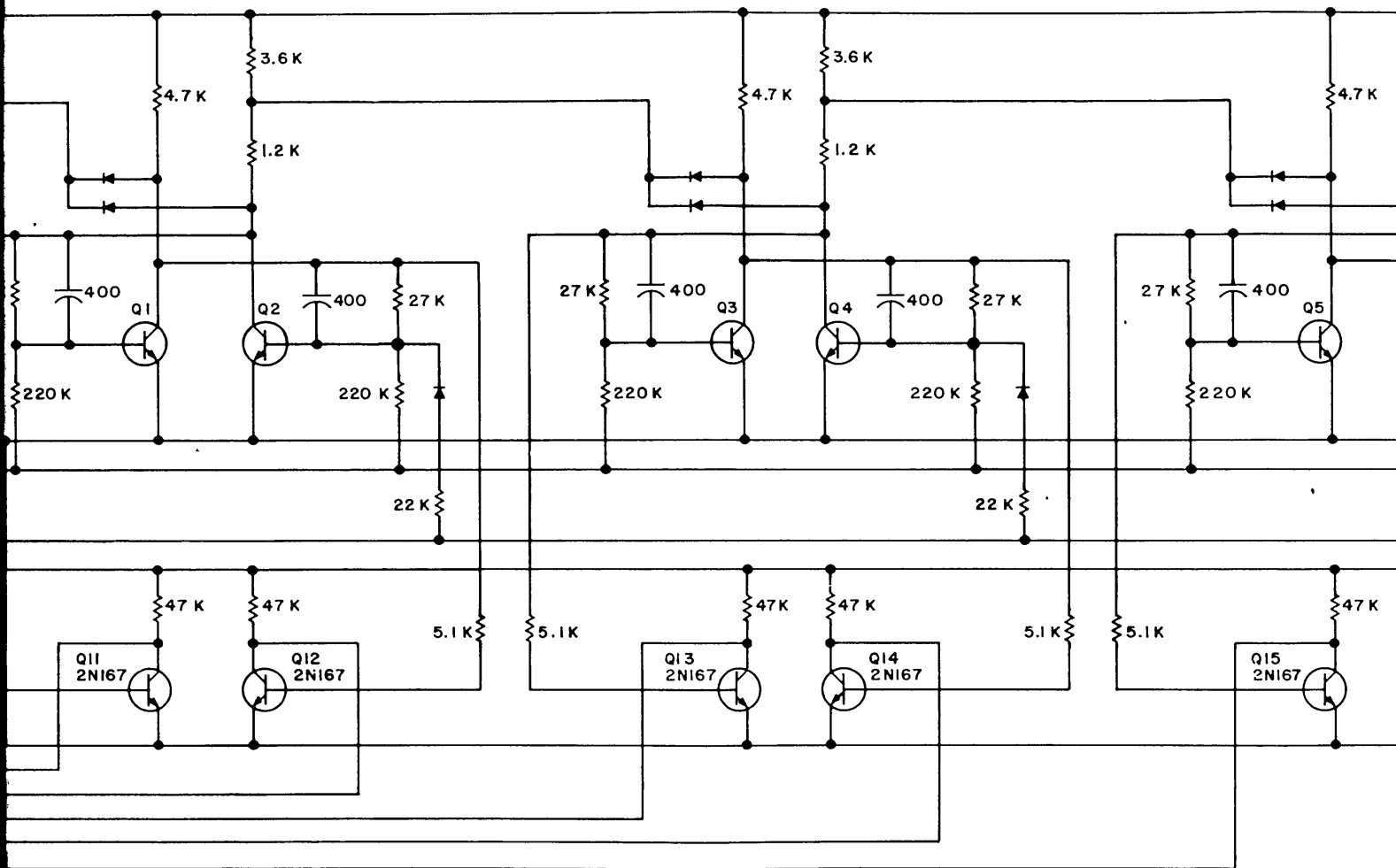
FIGURE 118. PARTIAL SCHEMATIC

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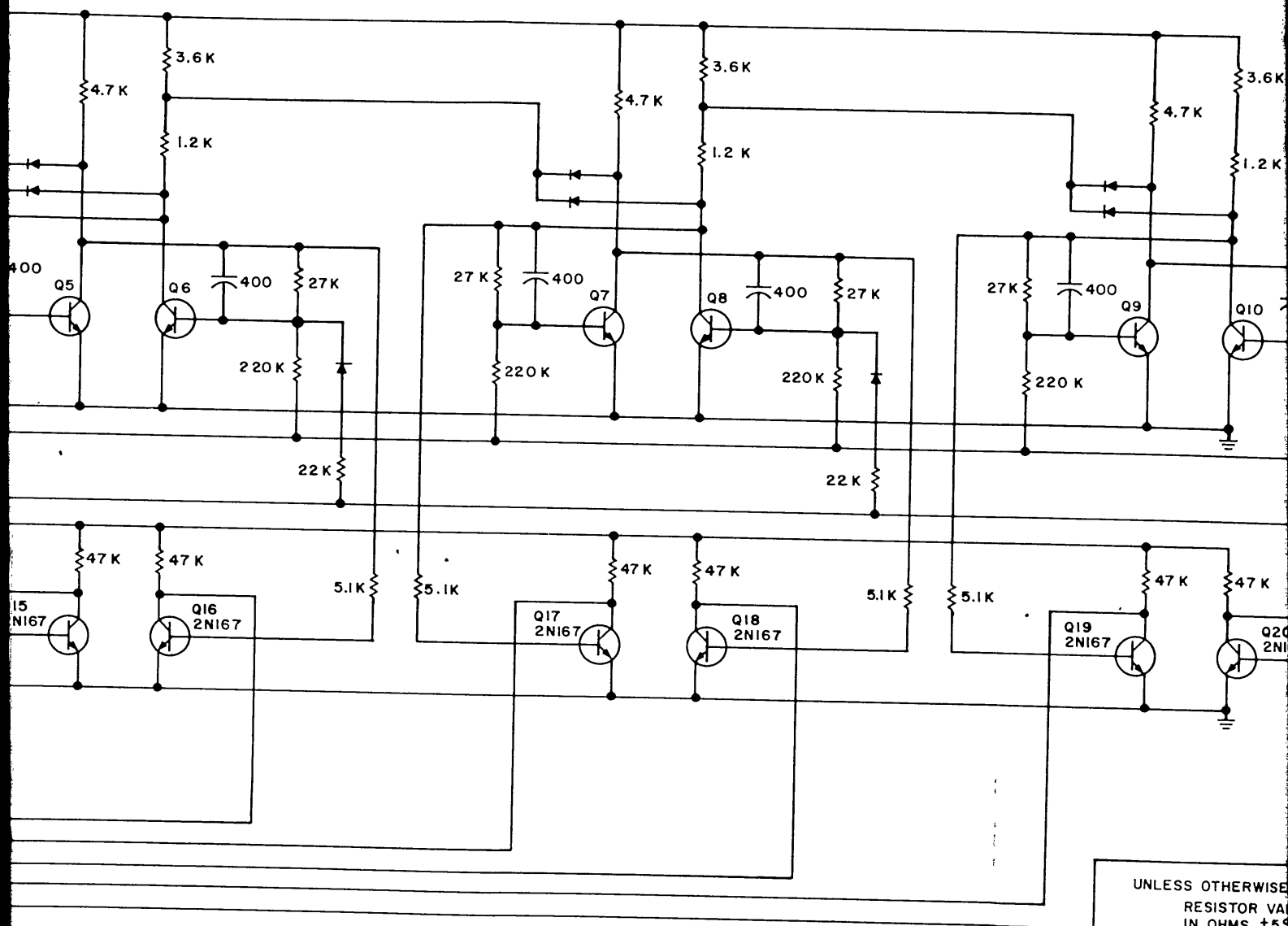


FIGURE II9. SCHEMATIC DIAGRAM OF TEST-PROB

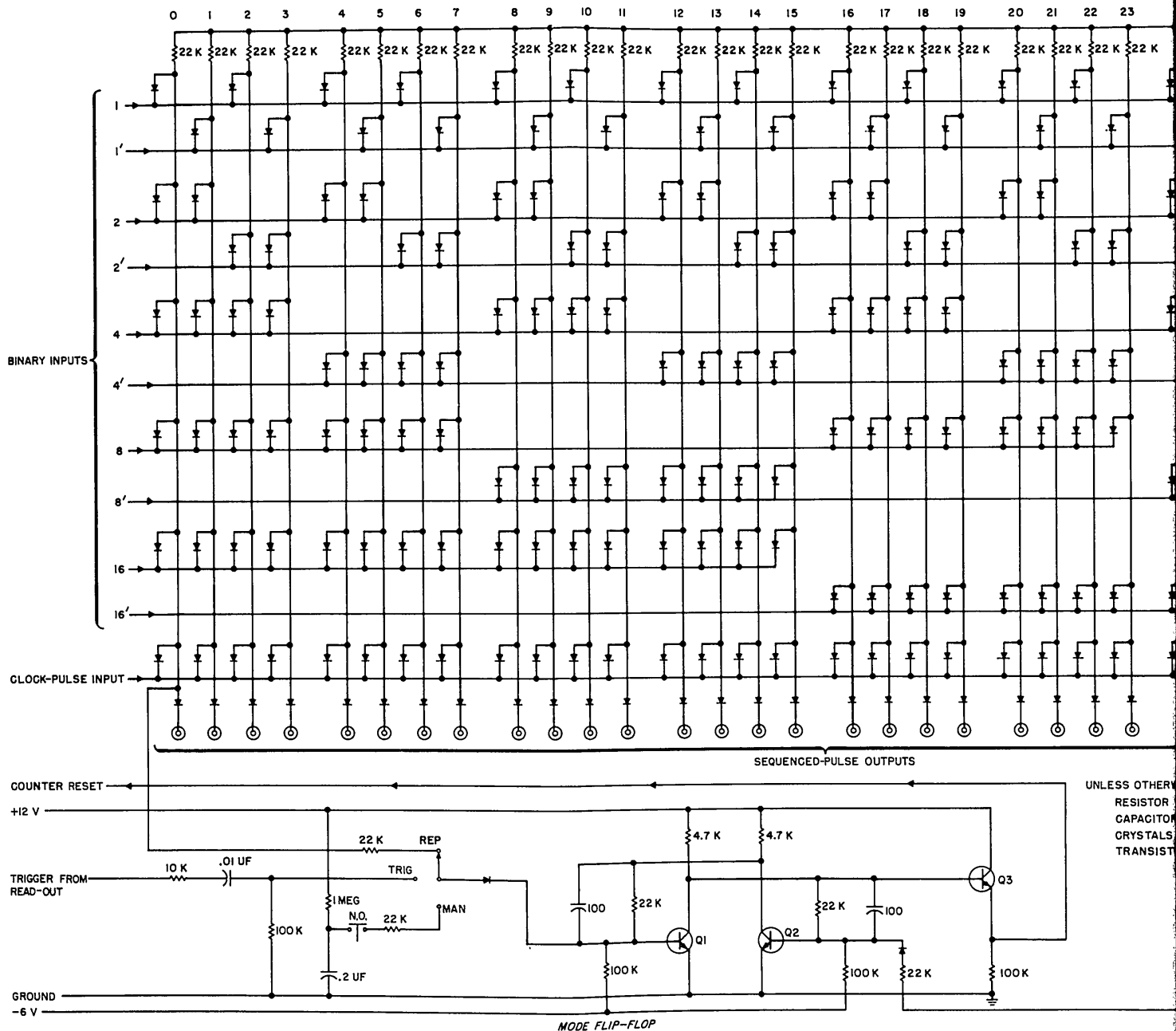
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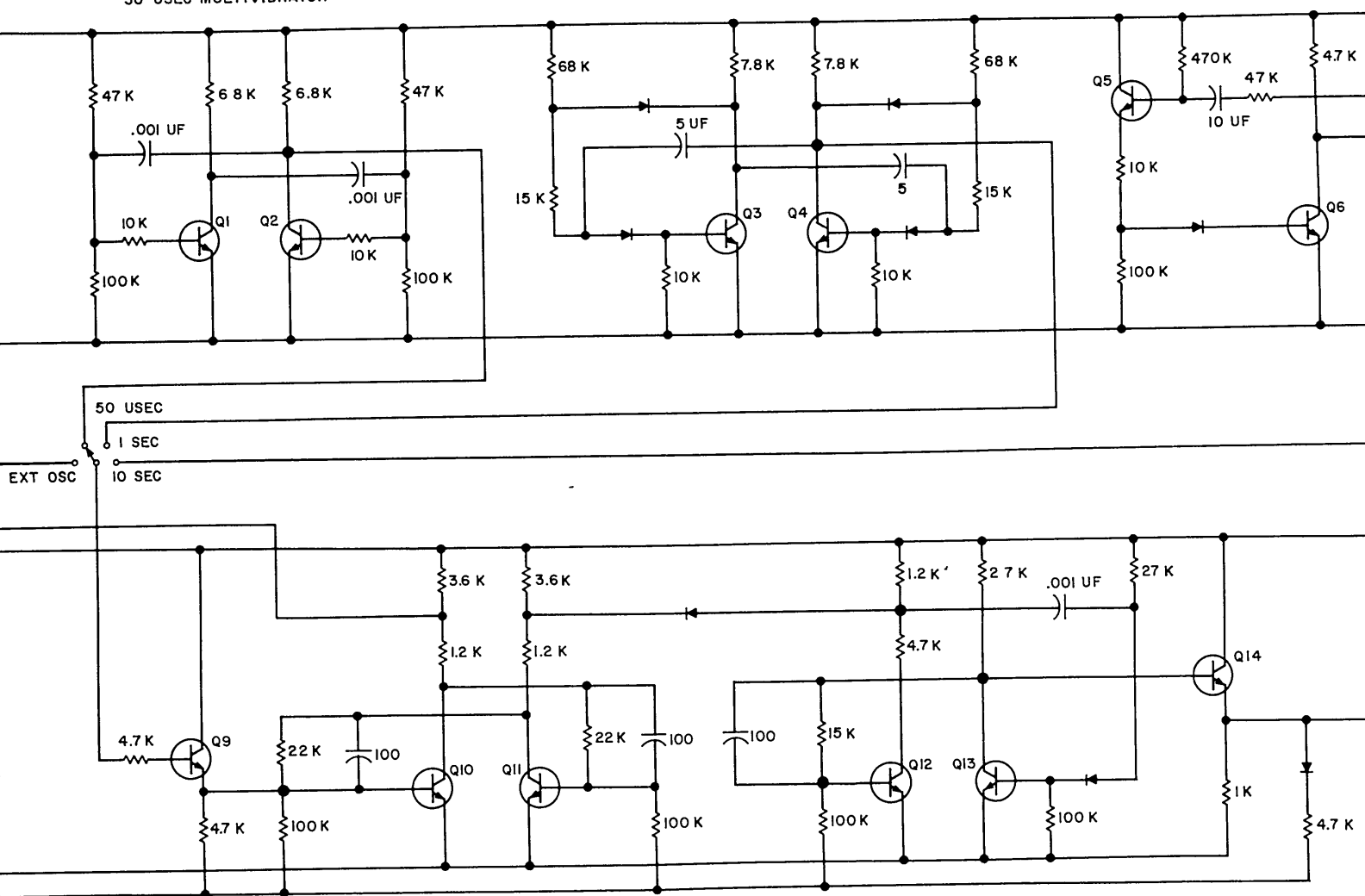
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CO

50-USEC MULTIVIBRATOR

1-SEC MULTIVIBRATOR

10-SEC MULTIVIBRATOR



UNLESS OTHERWISE NOTED:

RESISTOR VALUES ARE GIVEN IN OHMS $\pm 5\%$, $\frac{1}{2}$ WATT

CAPACITOR VALUES ARE GIVEN IN UUF

CRYSTALS ARE TYPE IN625

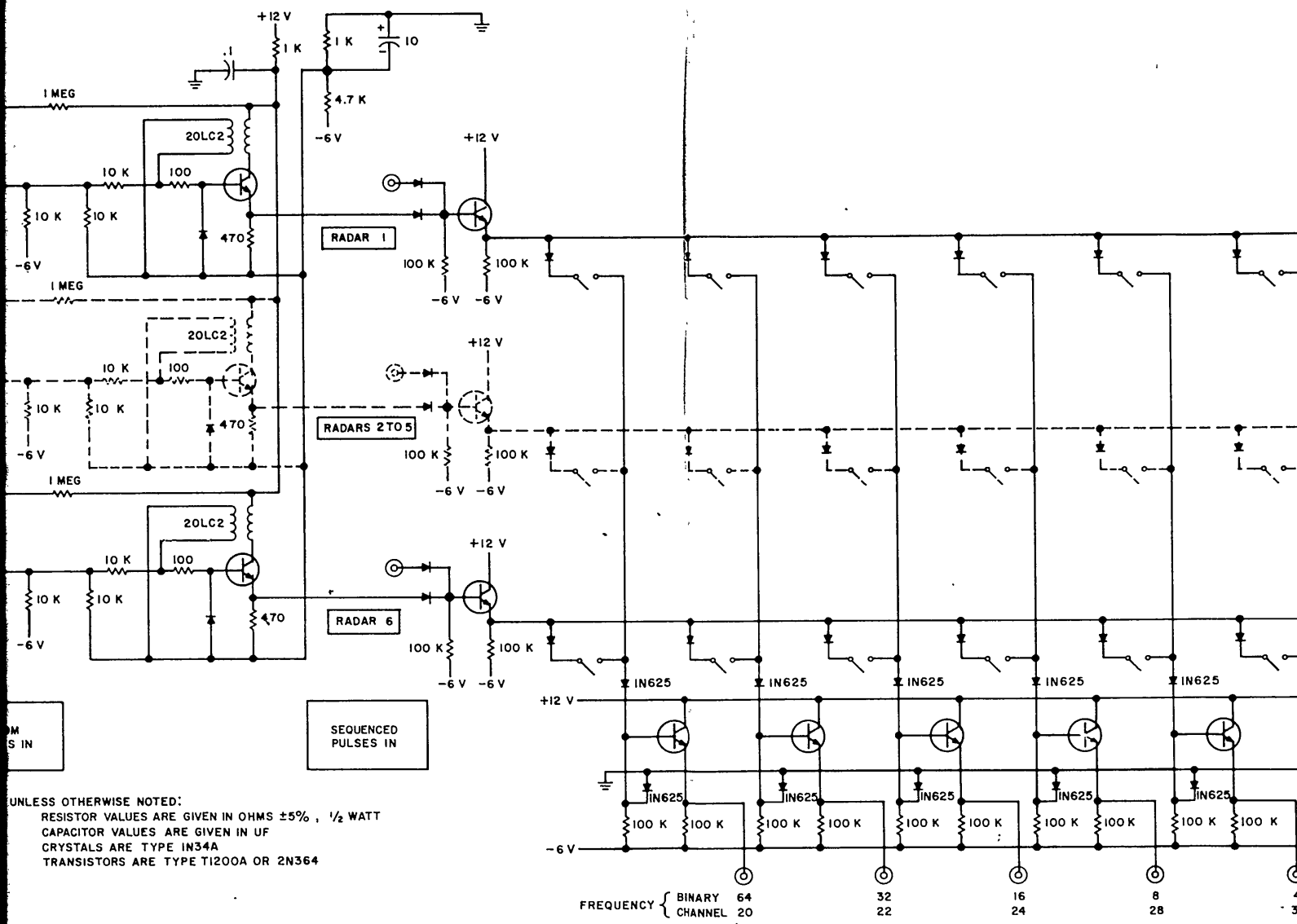
TRANSISTORS ARE TYPE 2N364

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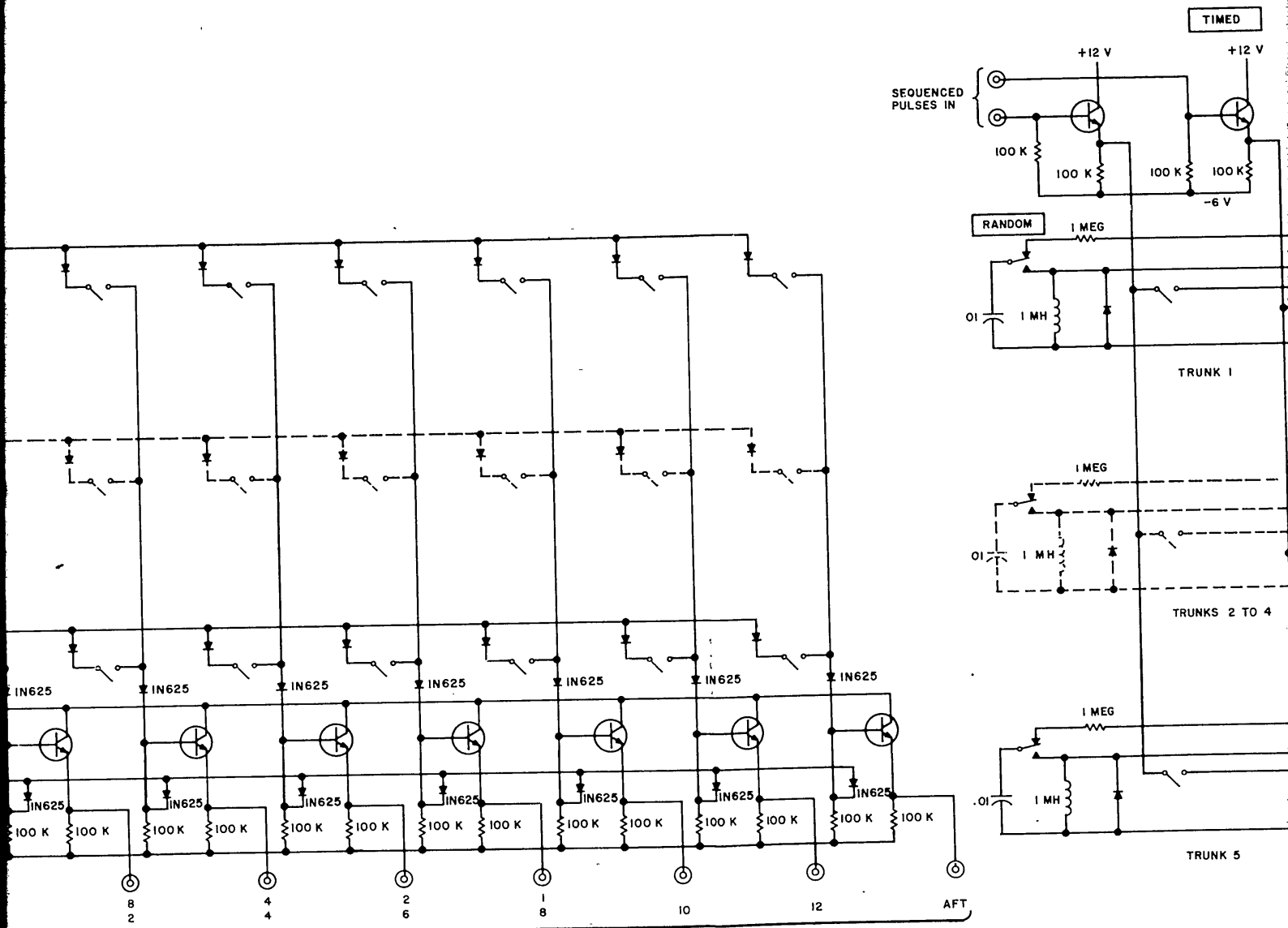


FIGURE 120. SCHEMATIC

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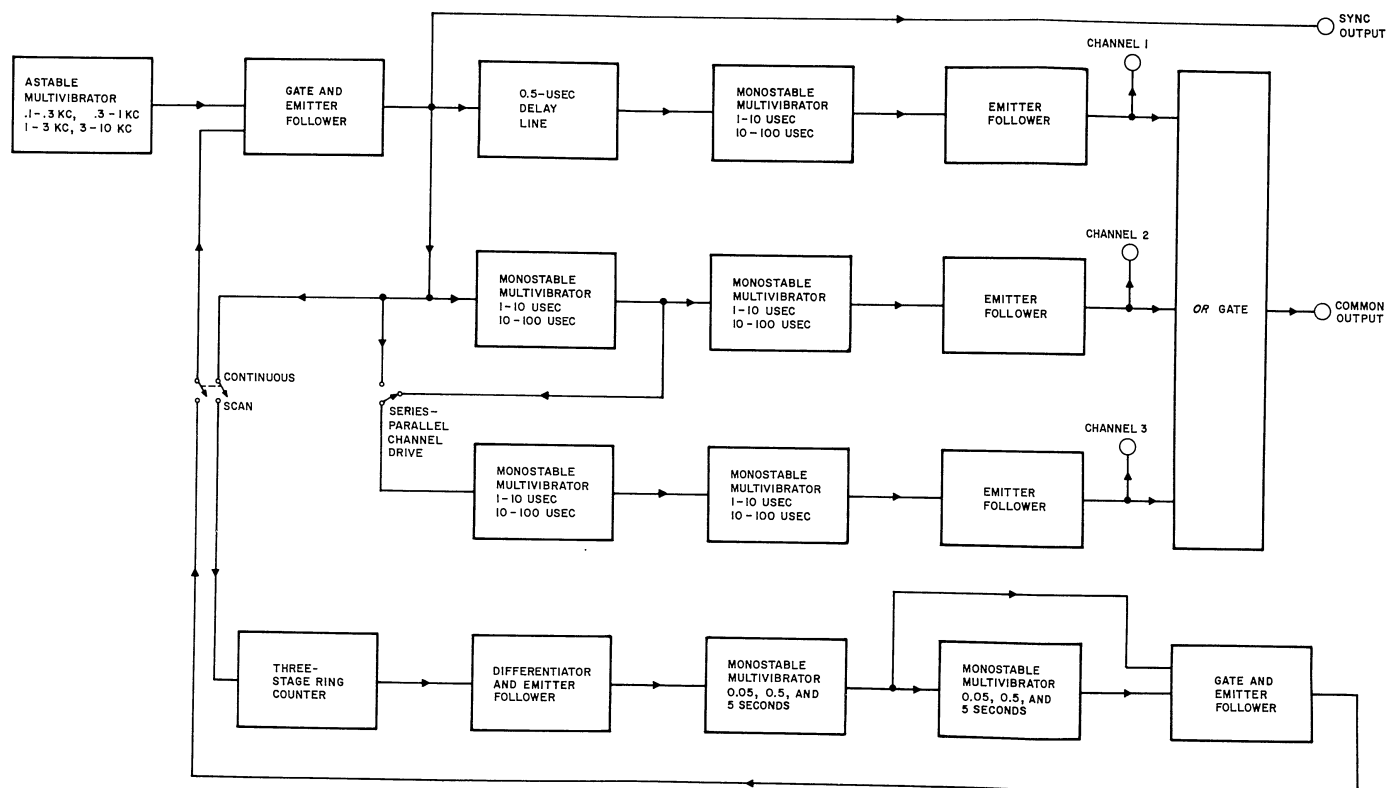


FIGURE 108. BLOCK DIAGRAM OF PULSE-TRAIN GENERATOR

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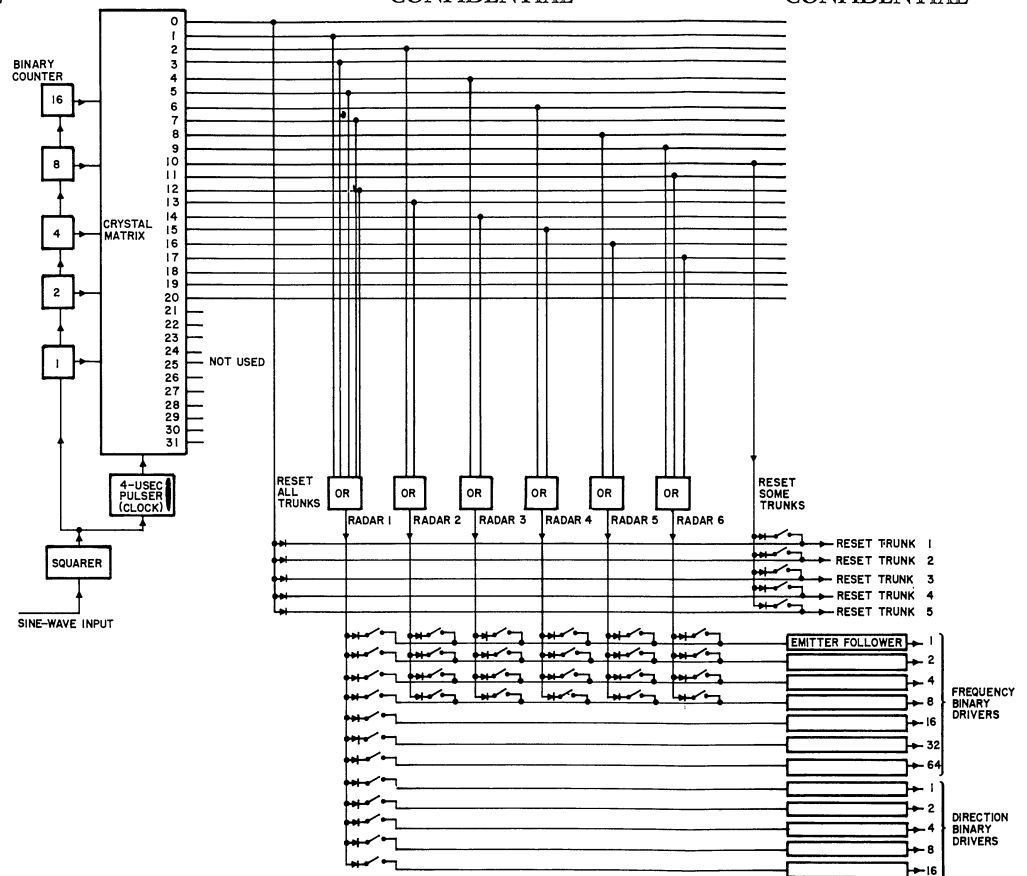


FIGURE 109. BLOCK DIAGRAM OF TEST-PROBLEM GENERATOR

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FIGURE 109
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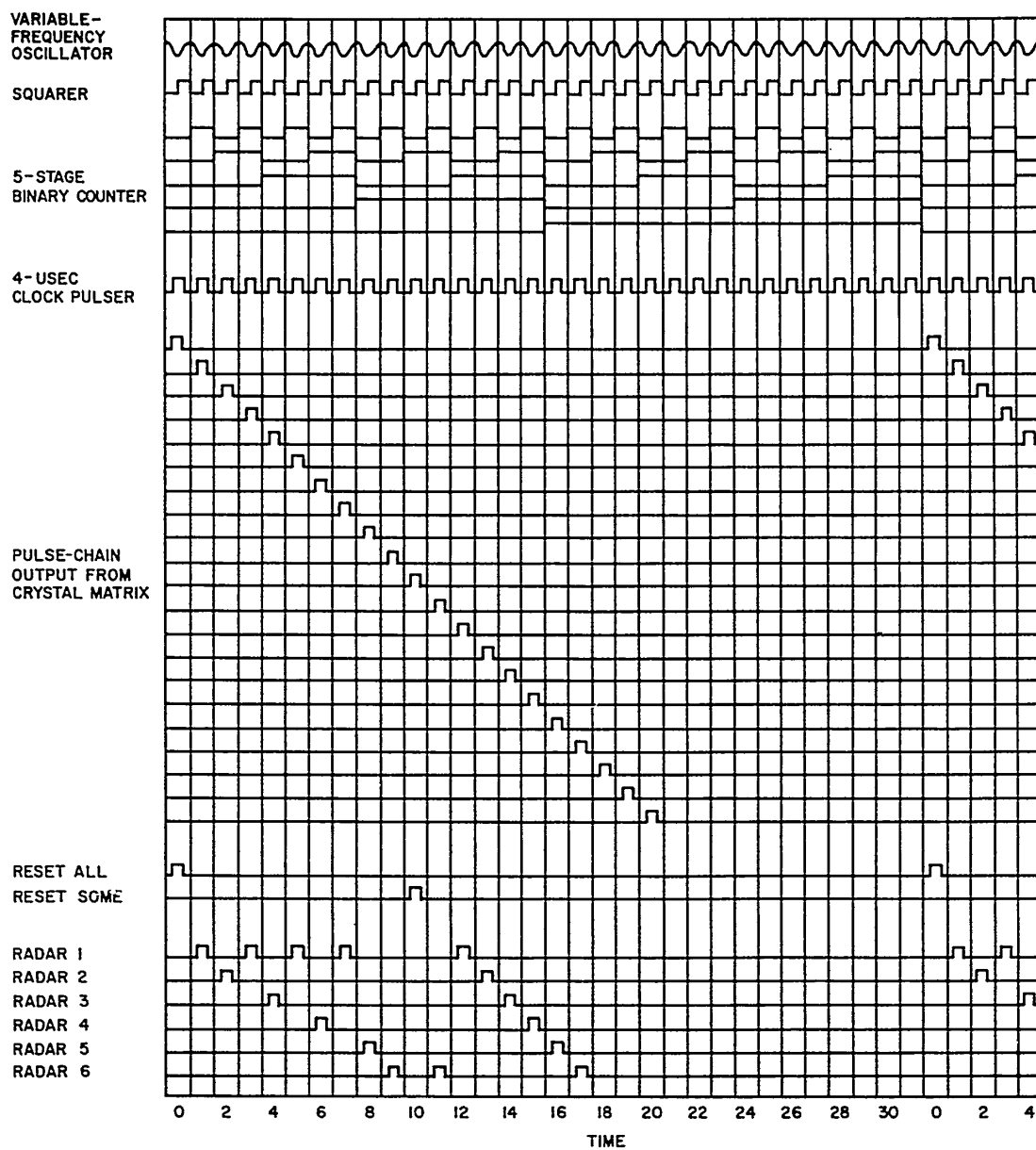
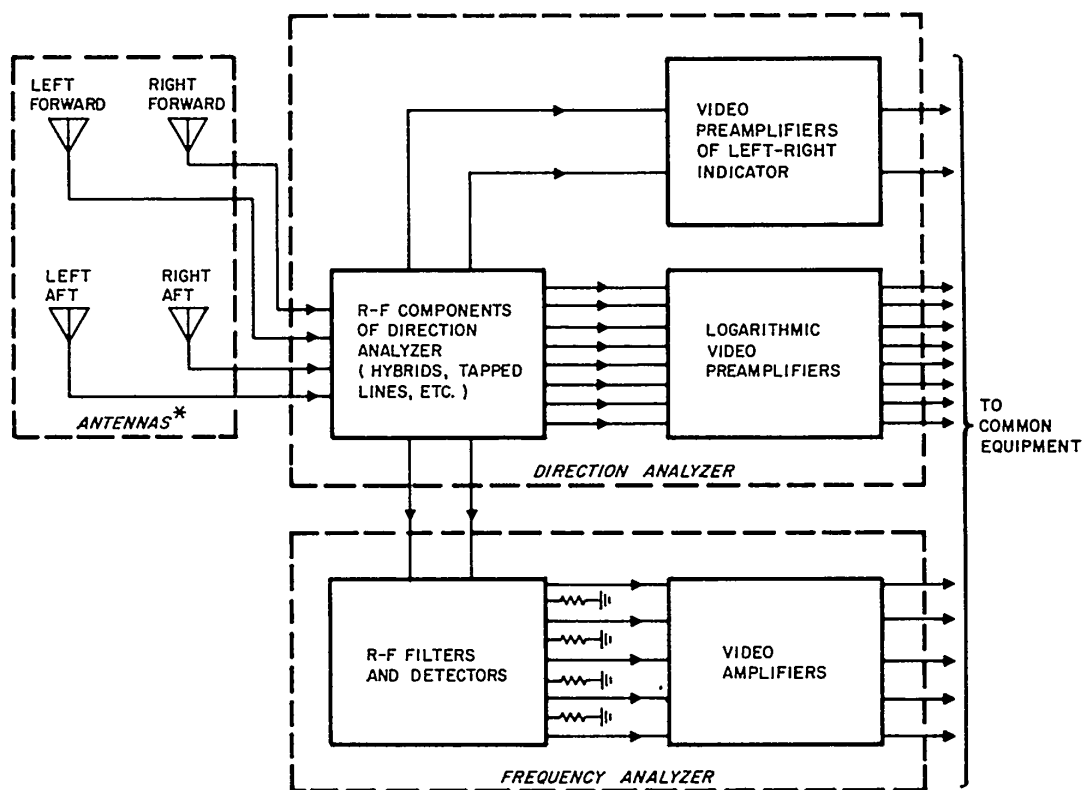


FIGURE IIO. TIMING DIAGRAM OF TEST-PROBLEM GENERATOR

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* ANTENNAS WERE 1/8-SCALE MODELS OF ANTENNAS FOR SUB-BAND 2, MOUNTED ON A 1/8-SCALE MODEL B-47, AND FULL-SCALE ANTENNAS FOR SUB-BAND 5 MOUNTED ON A MOCK-UP OF A SECTION OF A B-47.

FIGURE III. BLOCK DIAGRAM OF SUB-BAND 5 PORTION OF FEASIBILITY-DEMONSTRATION SYSTEM

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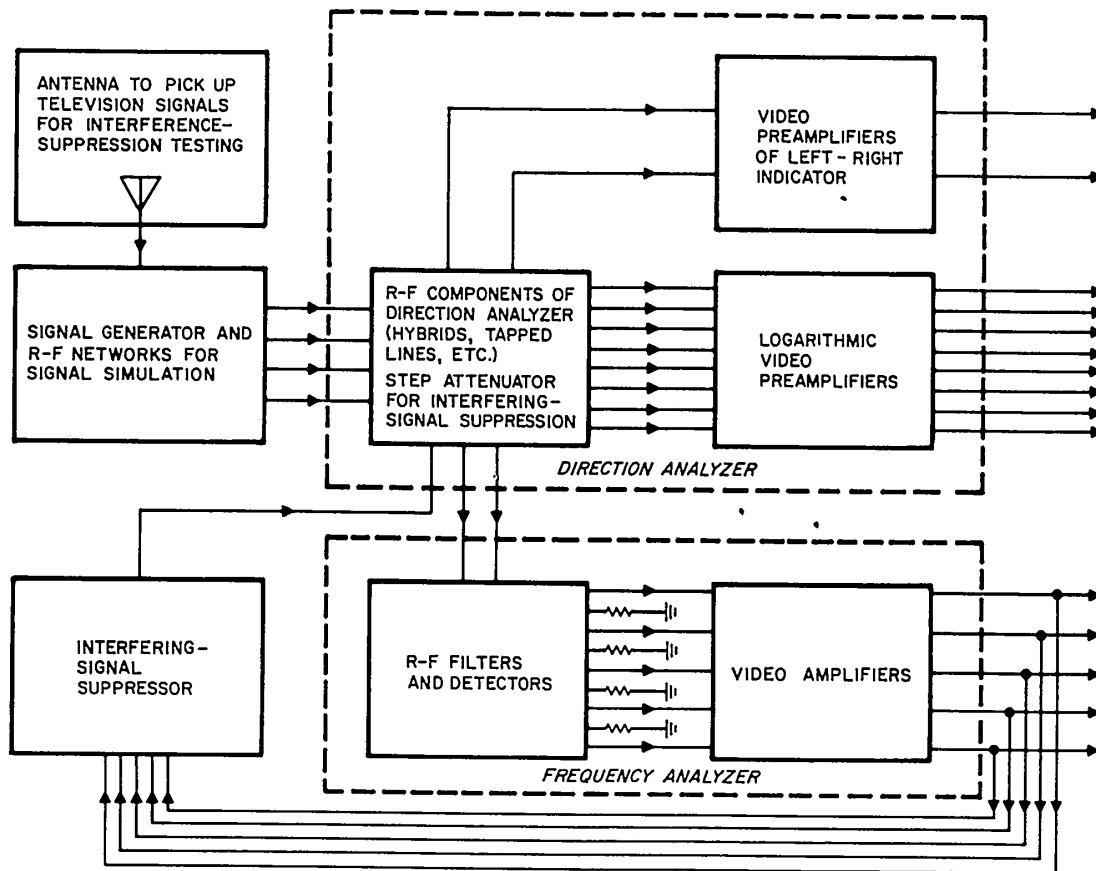


FIGURE II2. BLOCK DIAGRAM OF SUB-BAND 2 PORTION OF
FEASIBILITY-DEMONSTRATION SYSTEM

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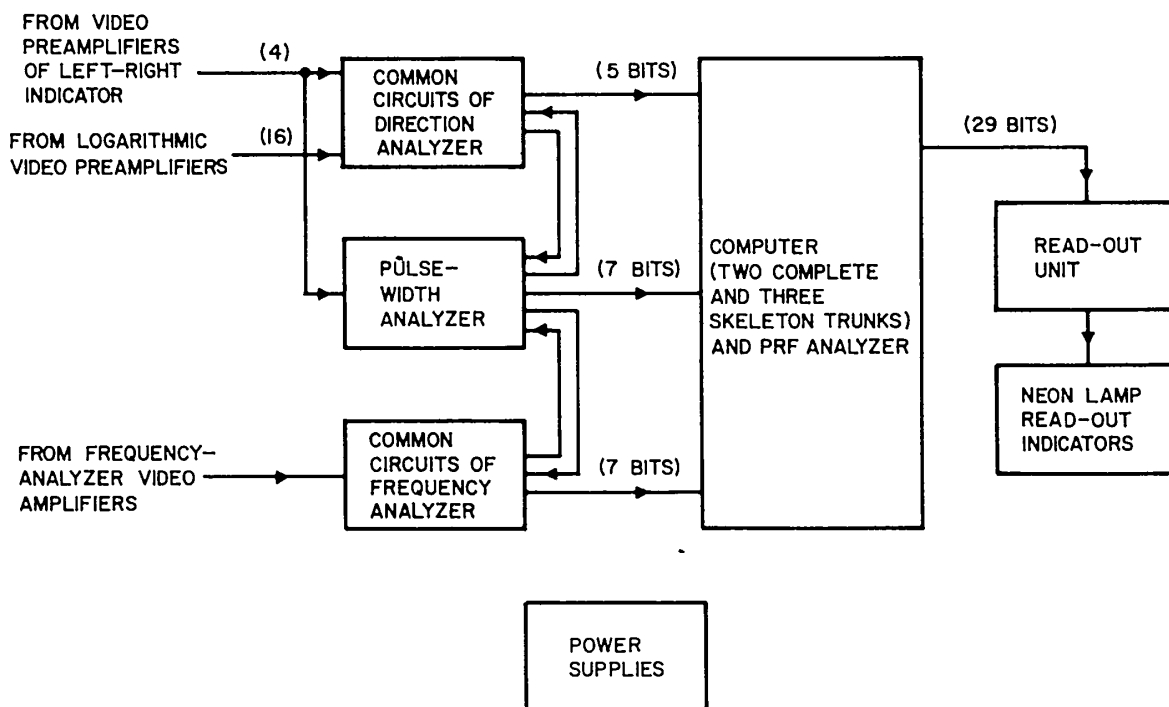


FIGURE I13. BLOCK DIAGRAM OF COMMON EQUIPMENT FOR FEASIBILITY-DEMONSTRATION SYSTEM

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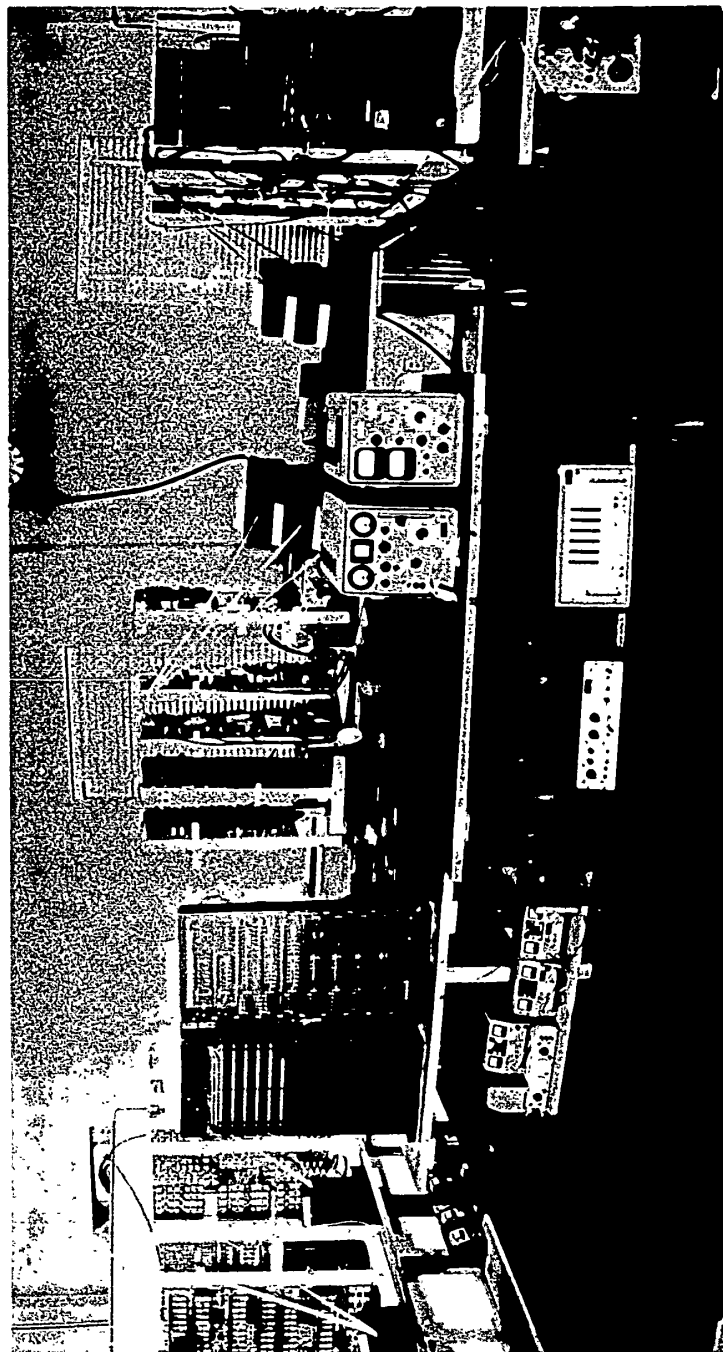


FIGURE 114. GENERAL VIEW OF FEASIBILITY-DEMONSTRATION SYSTEM

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AN/DLD-2
OUTPUT INDICATOR

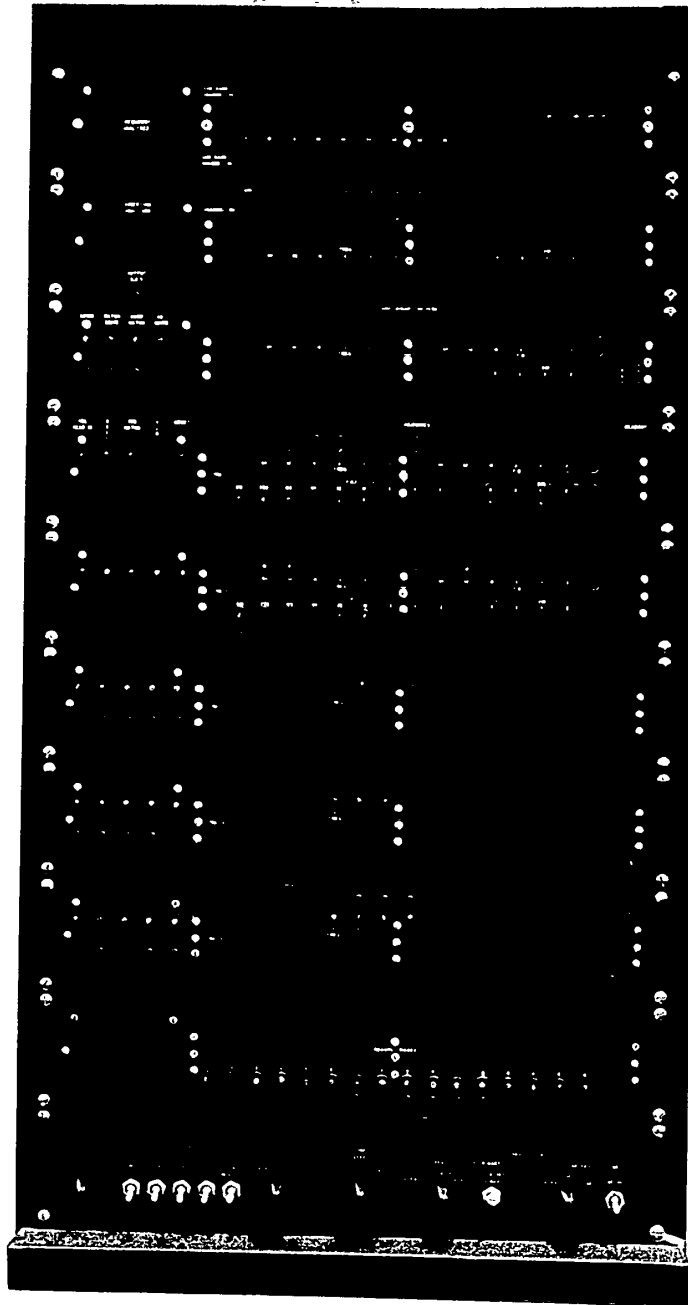


FIGURE 115. NEON-LAMP DISPLAY FOR READ-OUT OF FEASIBILITY-DEMONSTRATION SYSTEM

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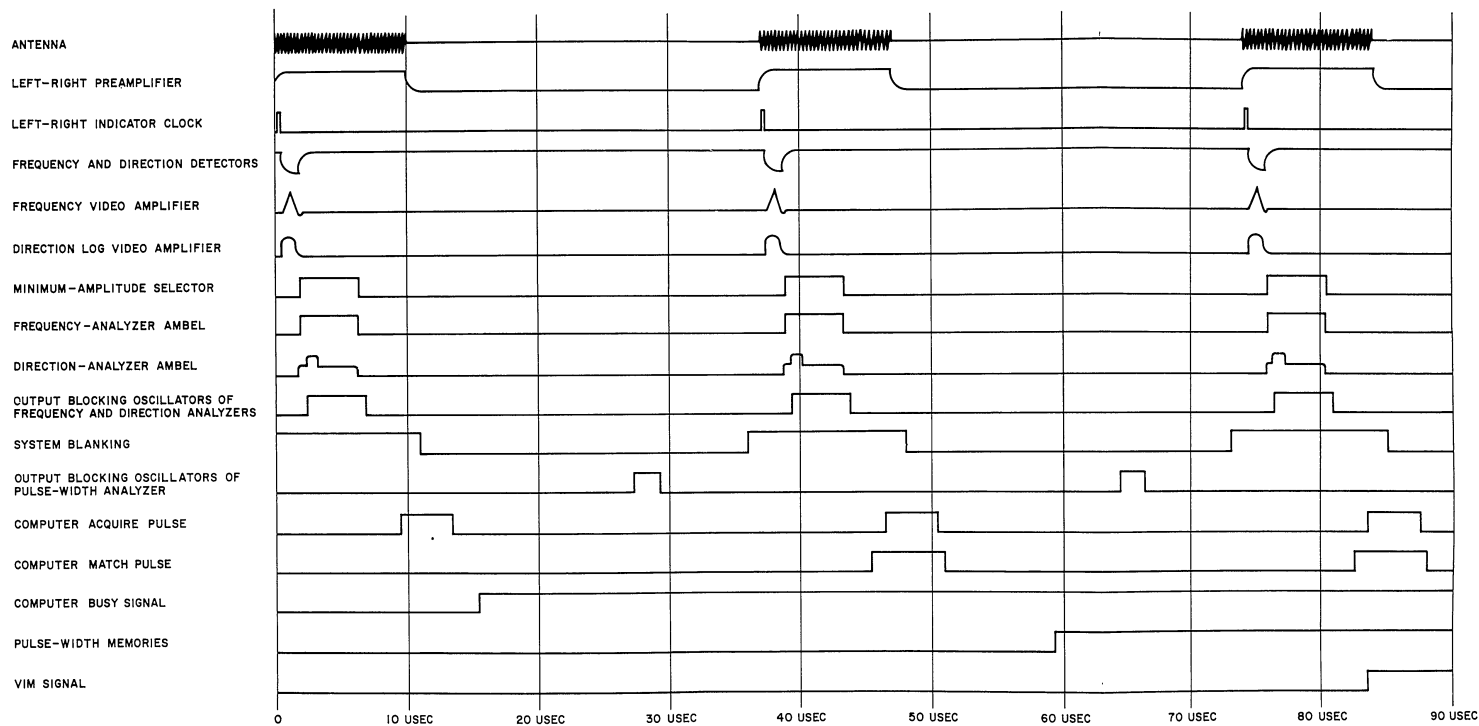


FIGURE 116. SYSTEM TIMING DIAGRAM

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FIGURE 116
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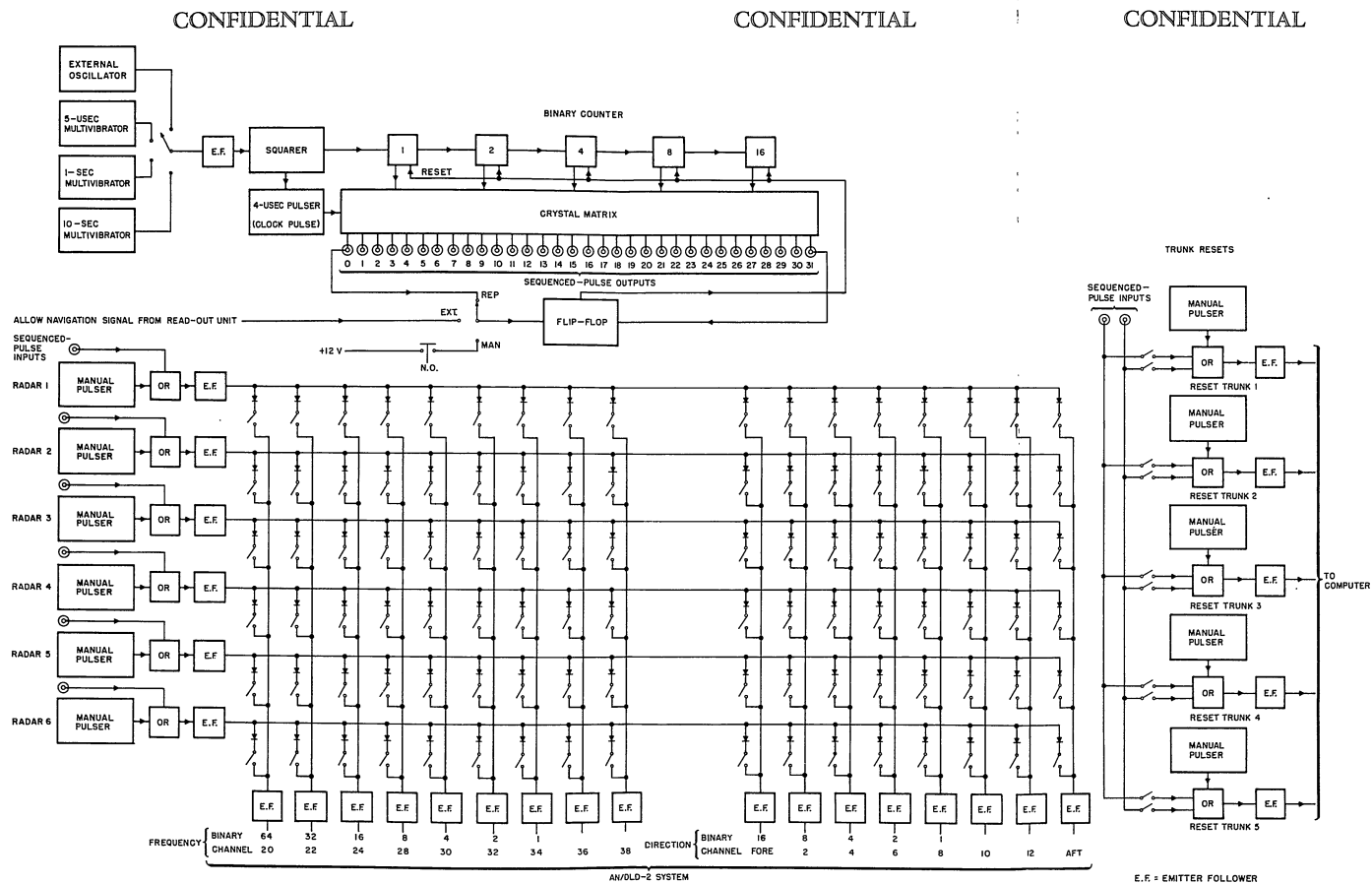
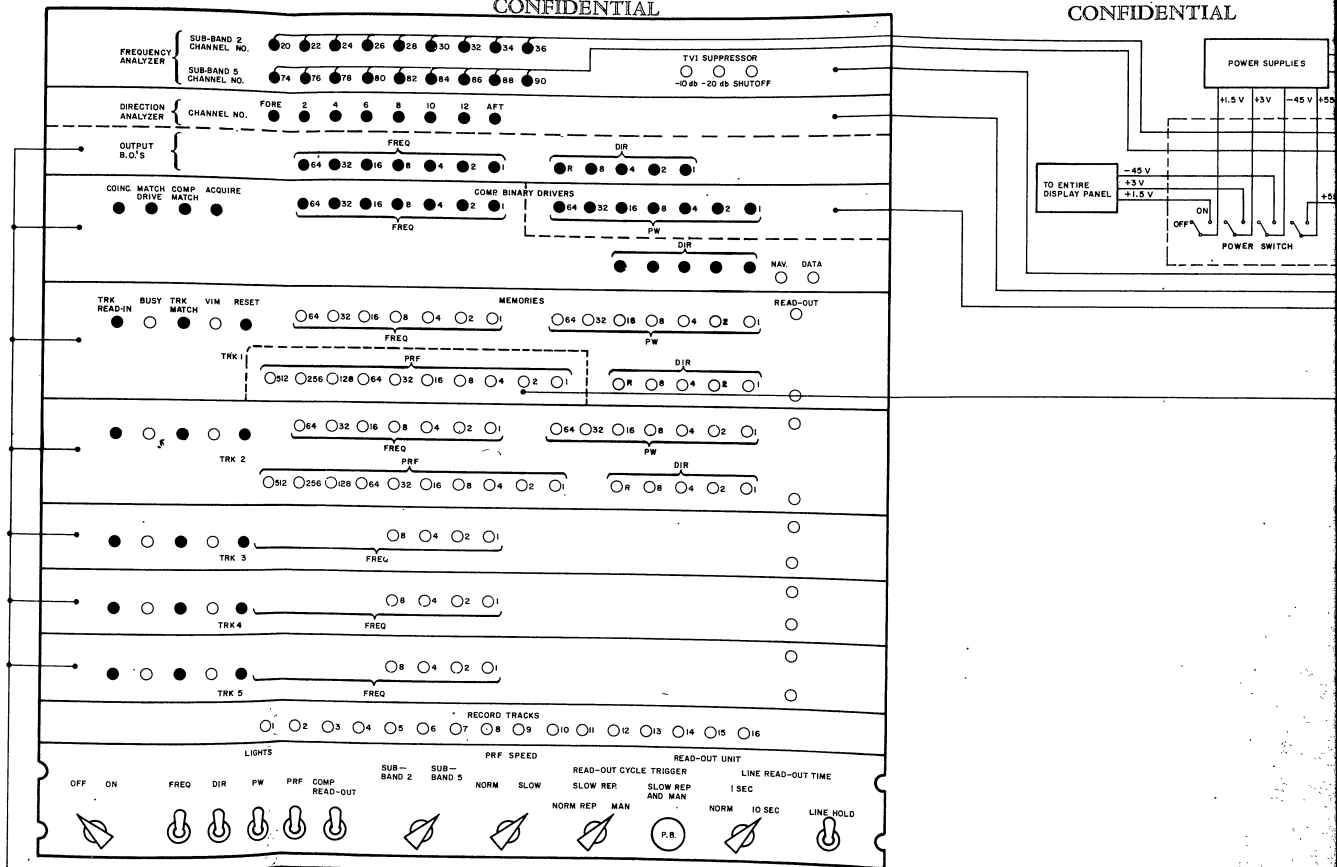


FIGURE 117. BLOCK DIAGRAM OF TEST-PROBLEM GENERATOR FOR FEASIBILITY-DEMONSTRATION SYSTEM

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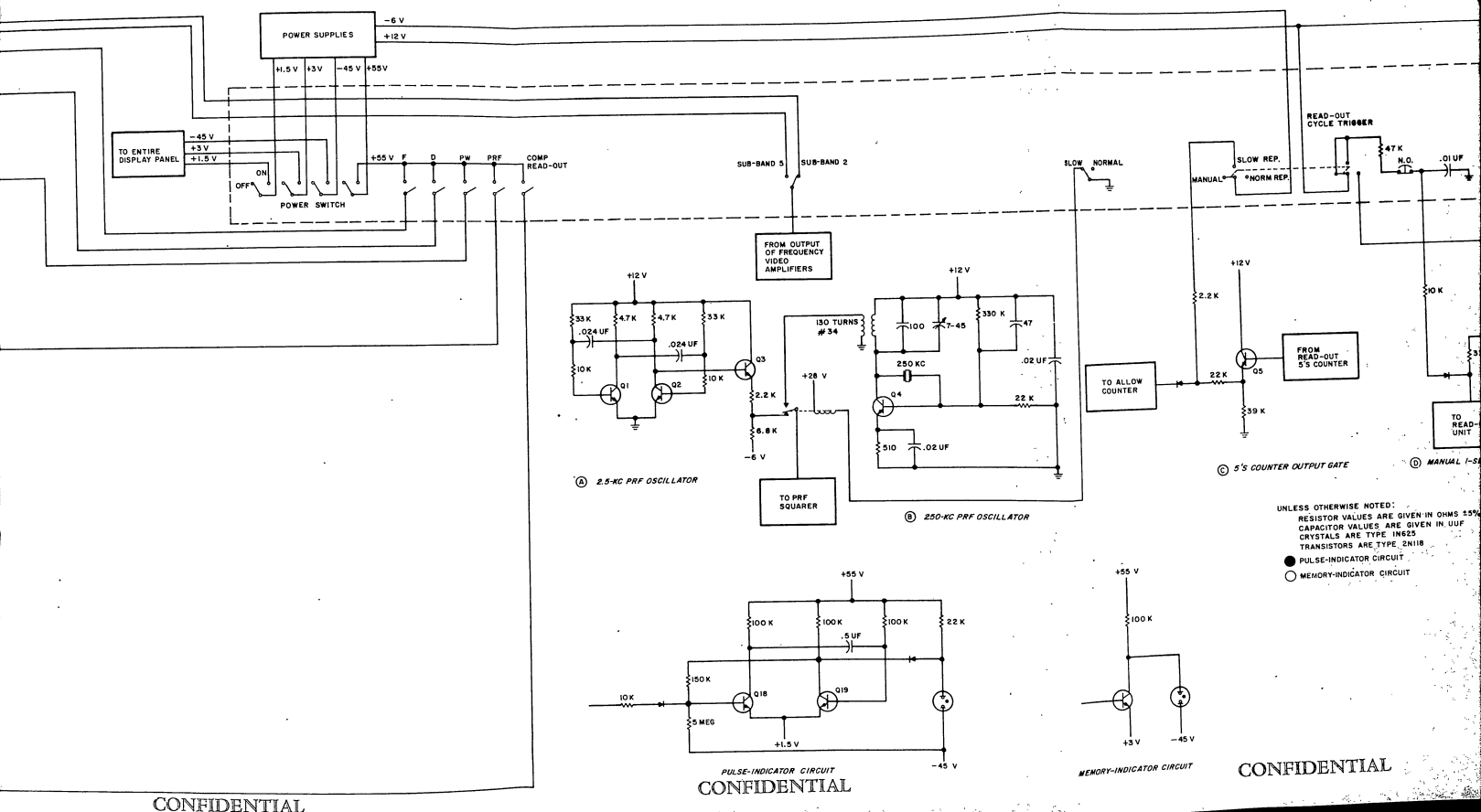
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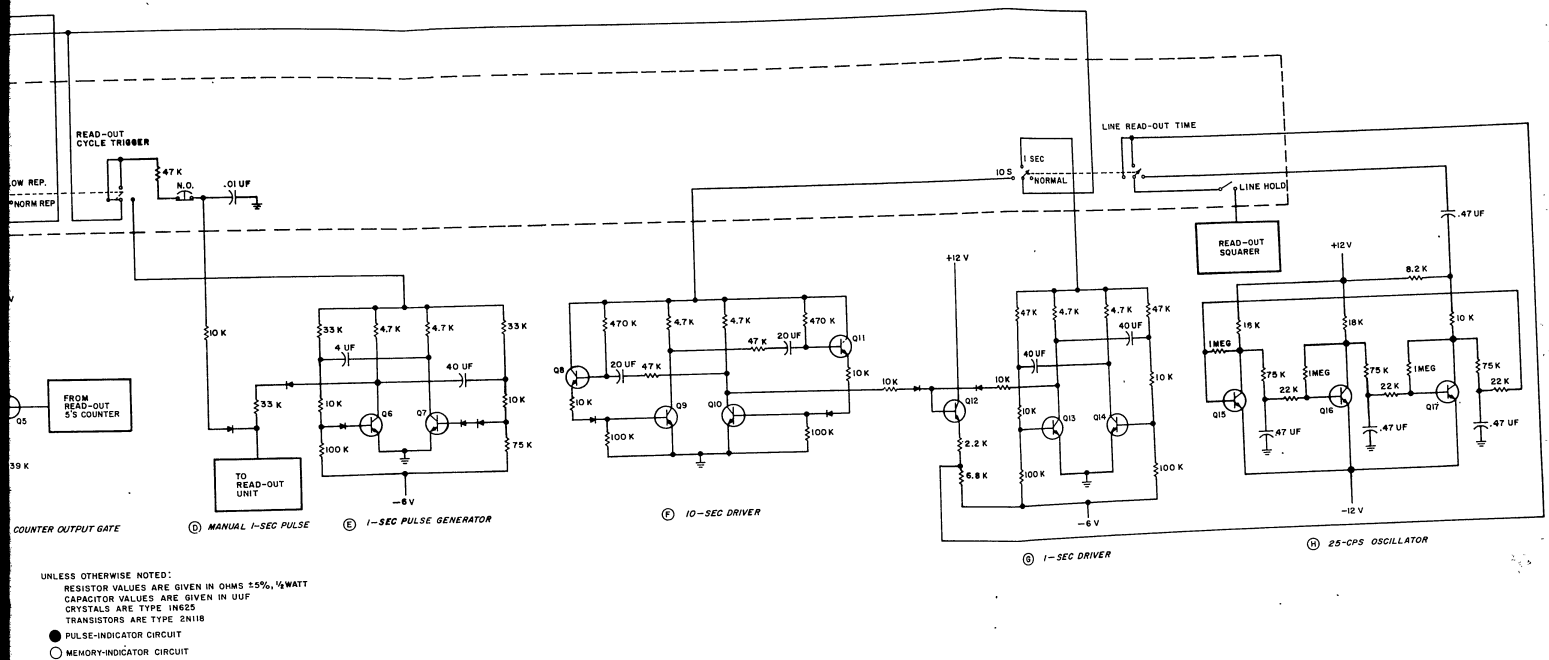


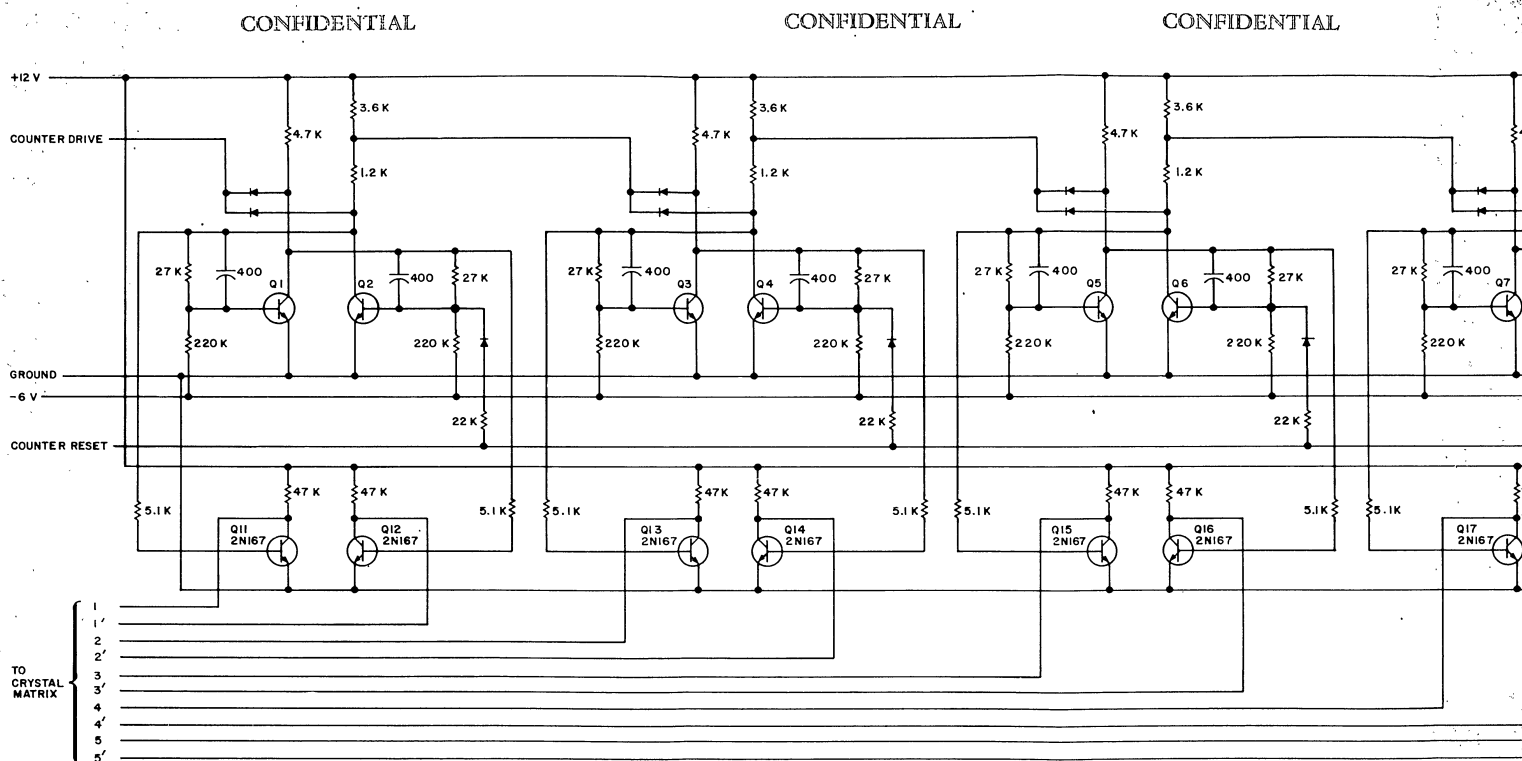
FIGURE 118. PARTIAL SCHEMATIC DIAGRAM OF NEON-LAMP DISPLAY

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FIGURE 118
377

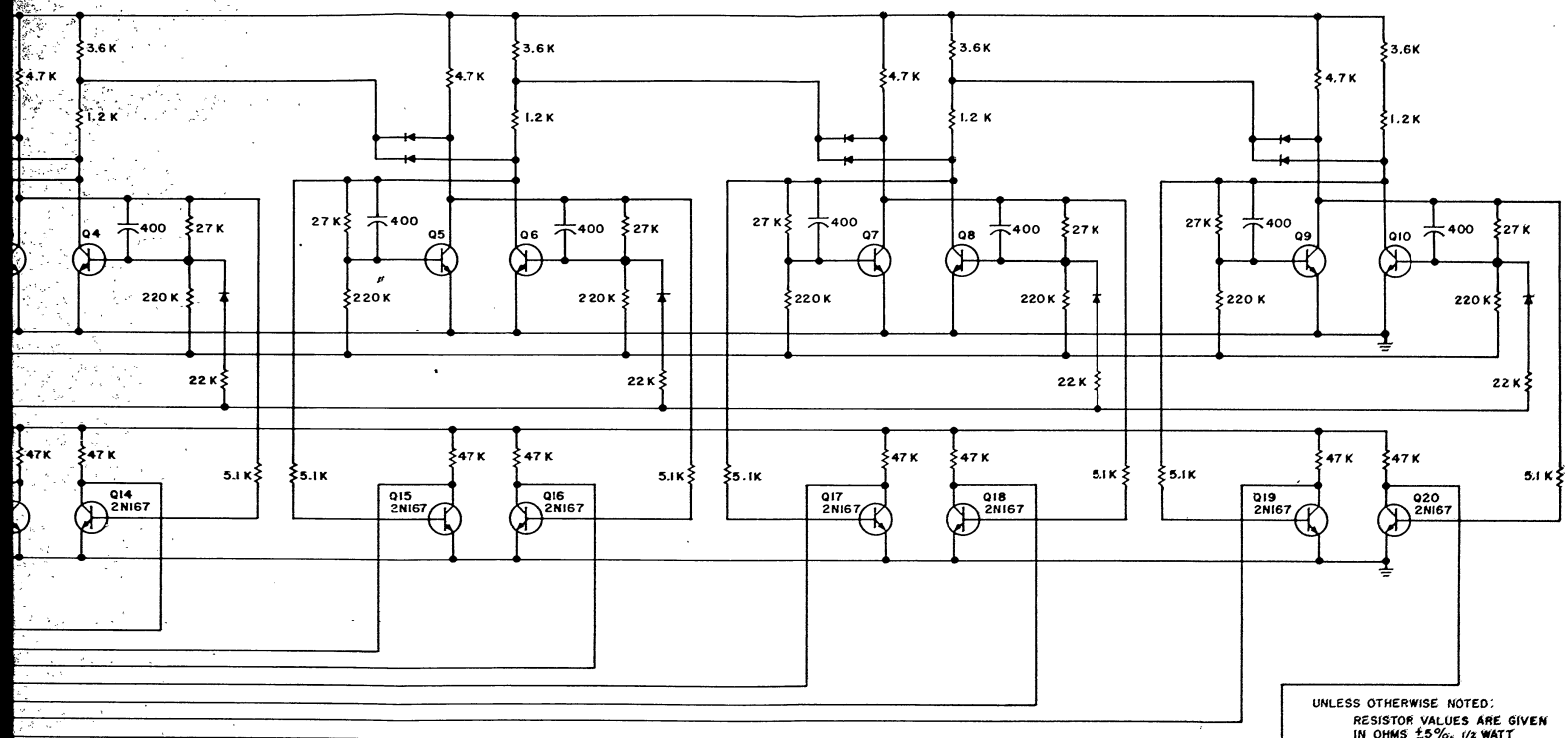


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A. BINARY COUNTER

FIGURE 119. SCHEMATIC DIAGRAM OF TEST-PROBLEM GENERATOR

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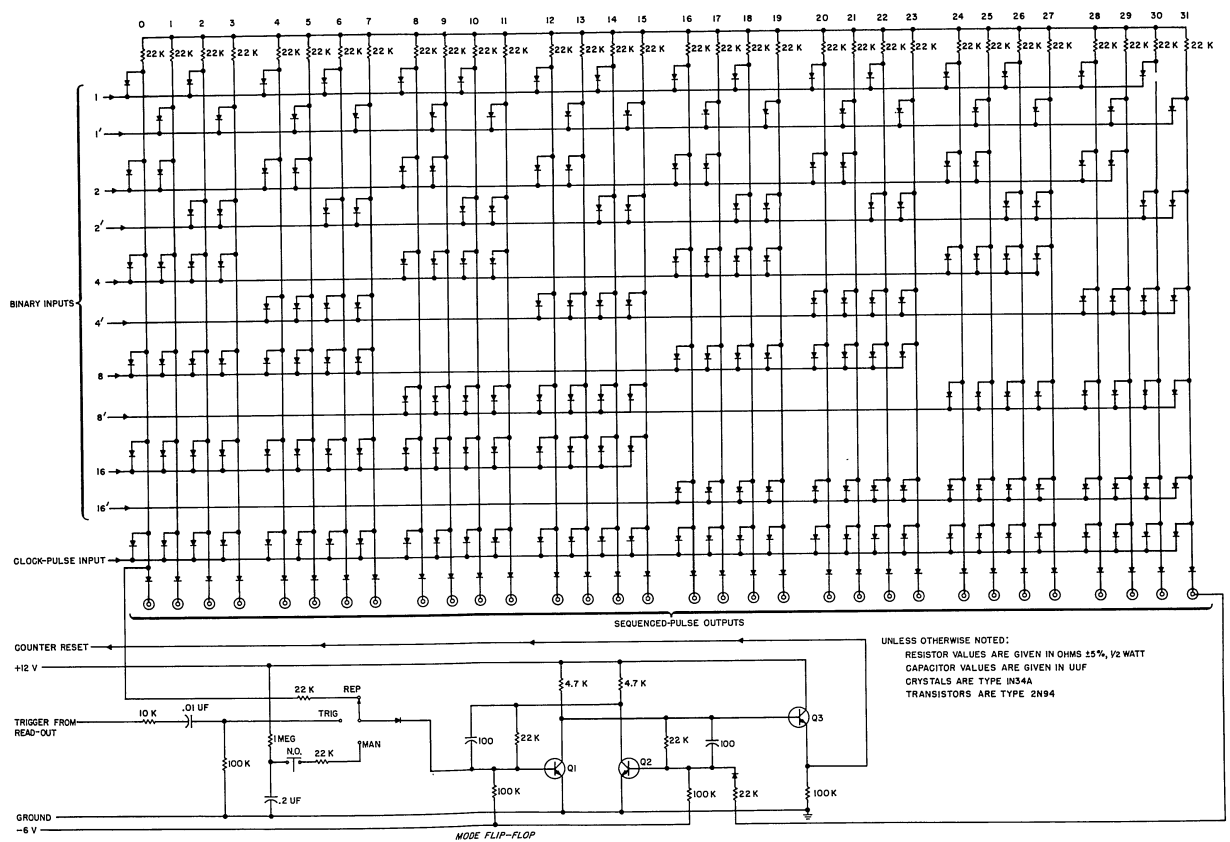
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FIGURE 119
SHEET 1 OF 3
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B. CRYSTAL MATRIX

FIGURE 119.

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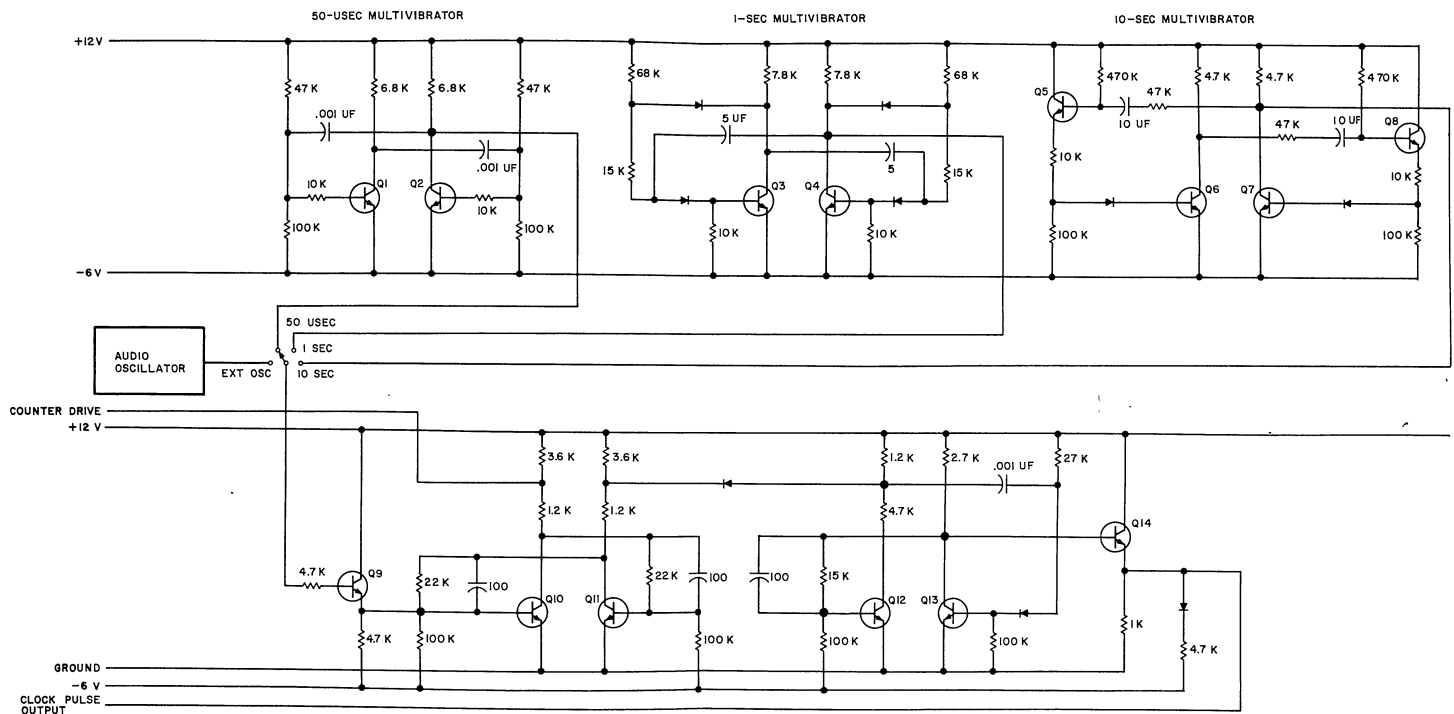
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SHEET 2 OF 3
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C. CLOCKS

FIGURE 119.

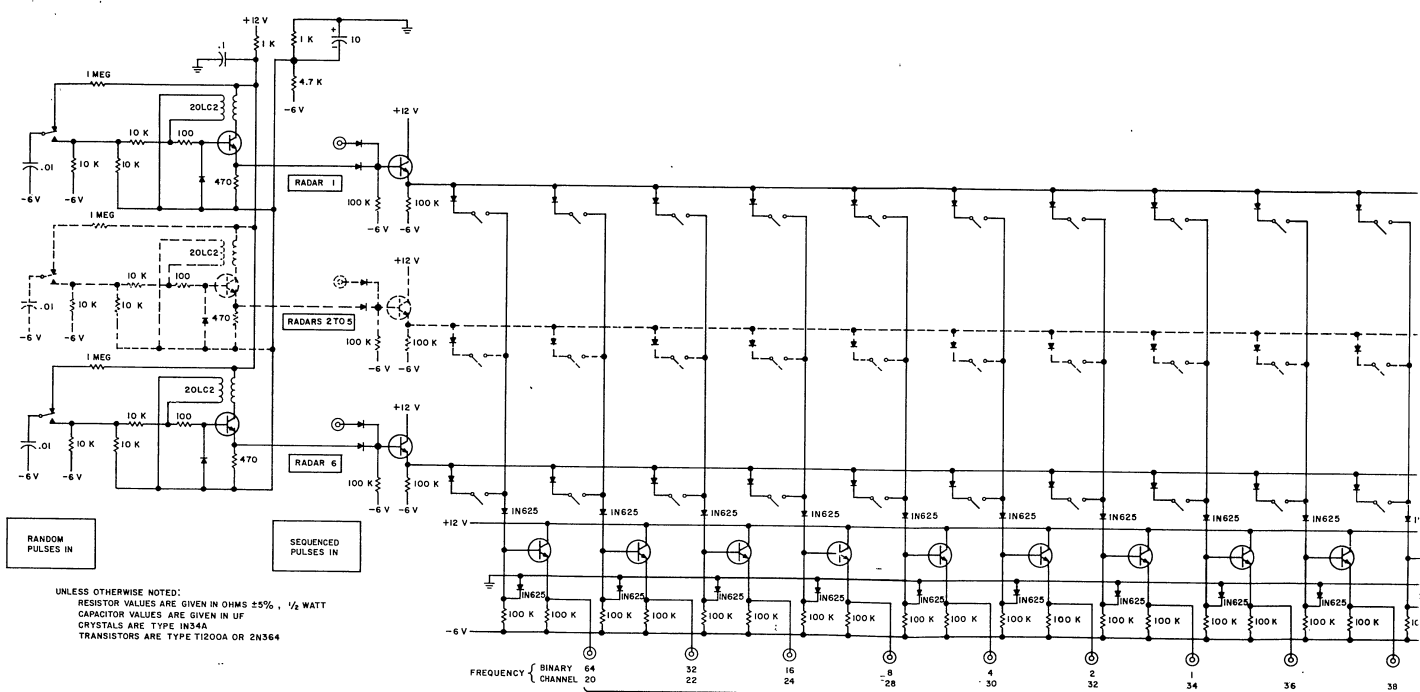
SHEET 3 OF 3

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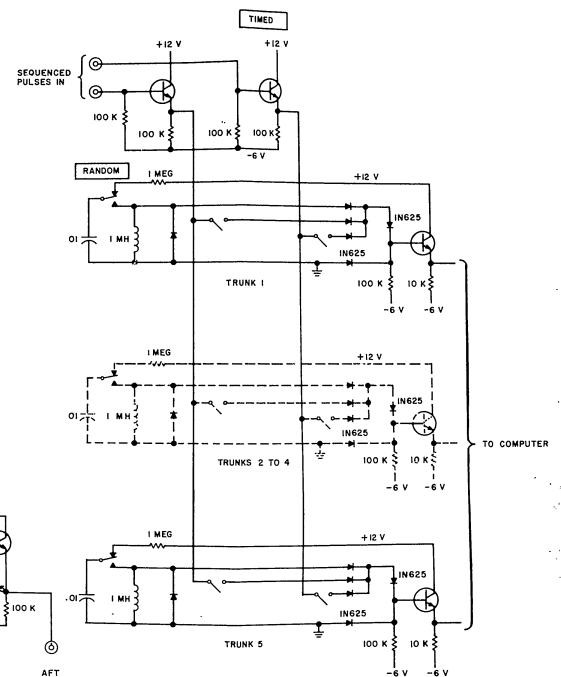
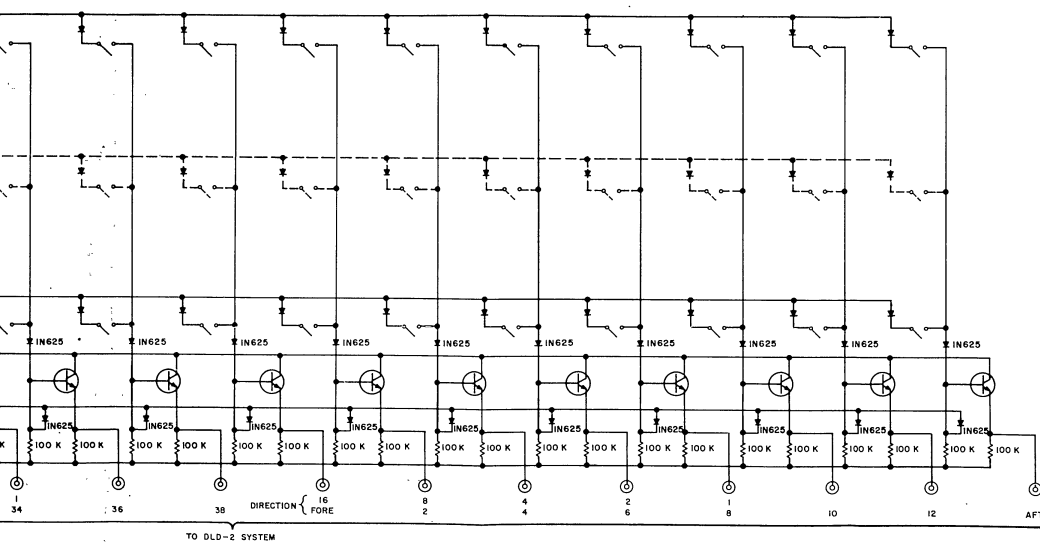


FIGURE 120. SCHEMATIC DIAGRAM OF TEST-PROBLEM CONTROLS

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FIGURE 120
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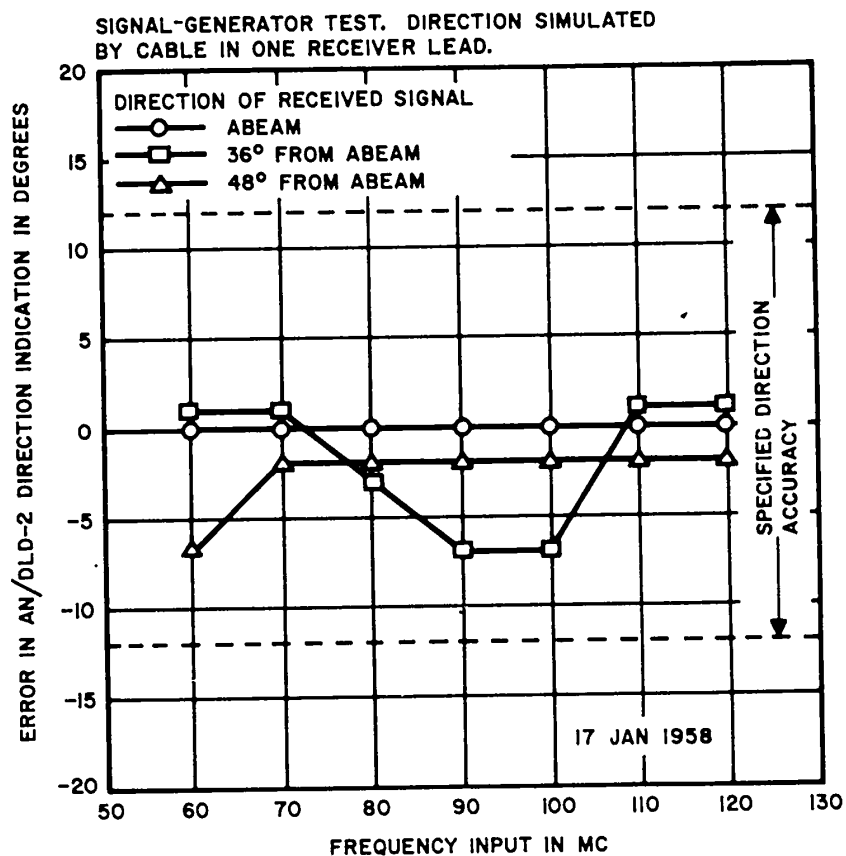
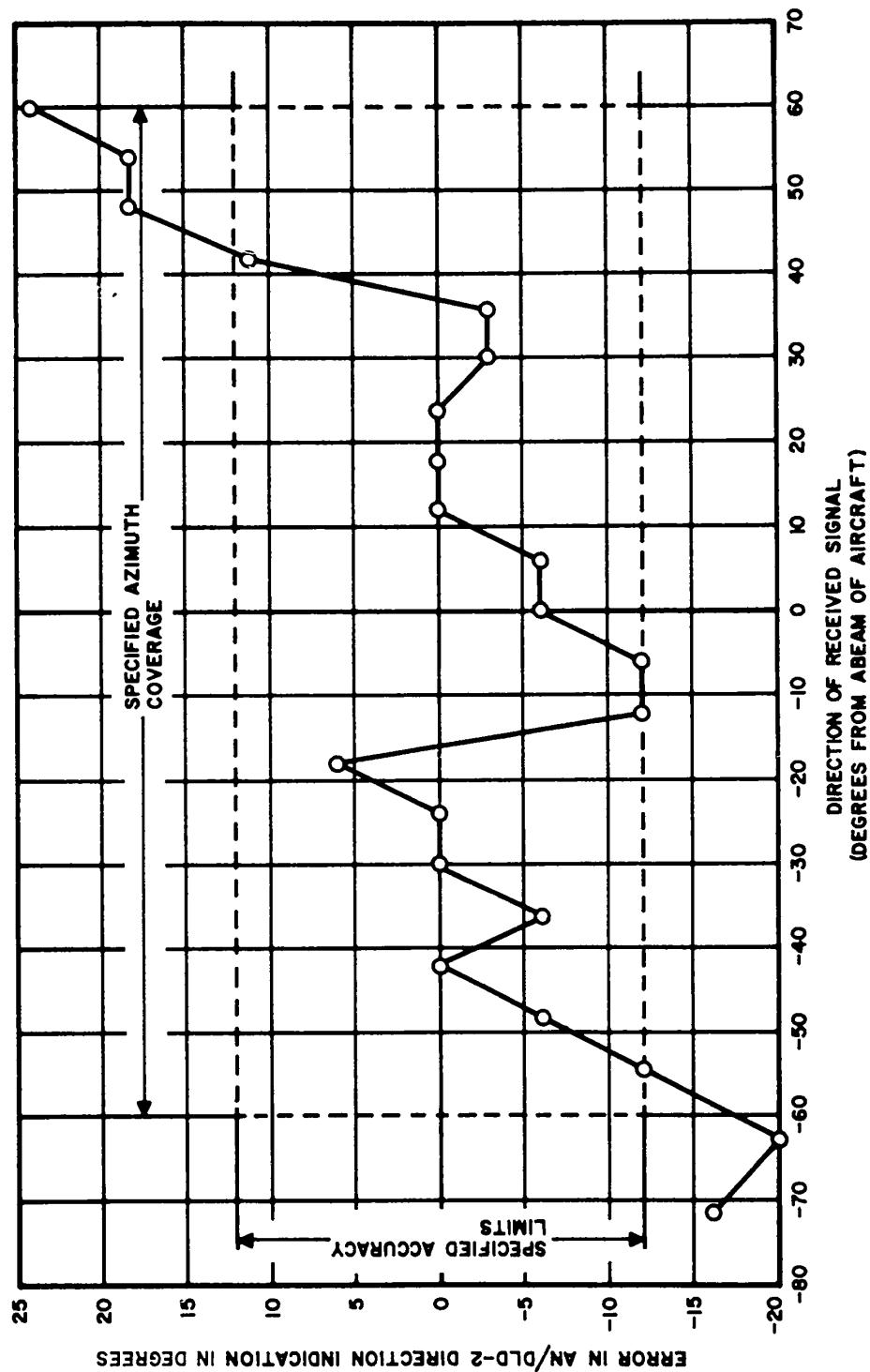


FIGURE 121. BEARING ERROR OF FEASIBILITY-DEMONSTRATION SYSTEM VS FREQUENCY IN SUB-BAND 2

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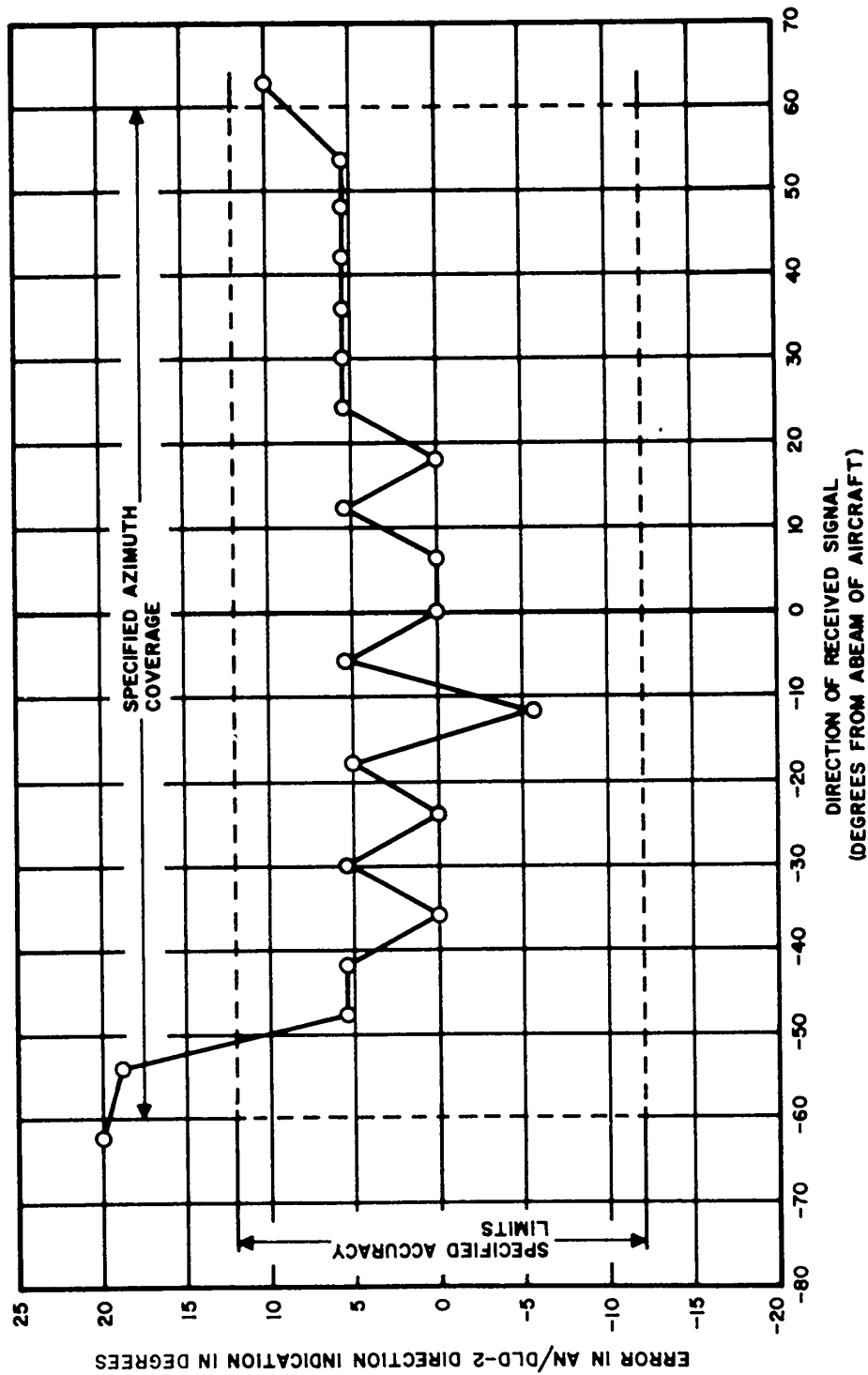


A. 69.75 MC SCALED TO 550 MC

FIGURE 122. BEARING ERROR OF FEASIBILITY-DEMONSTRATION SYSTEM VS BEARING IN SUB-BAND 2

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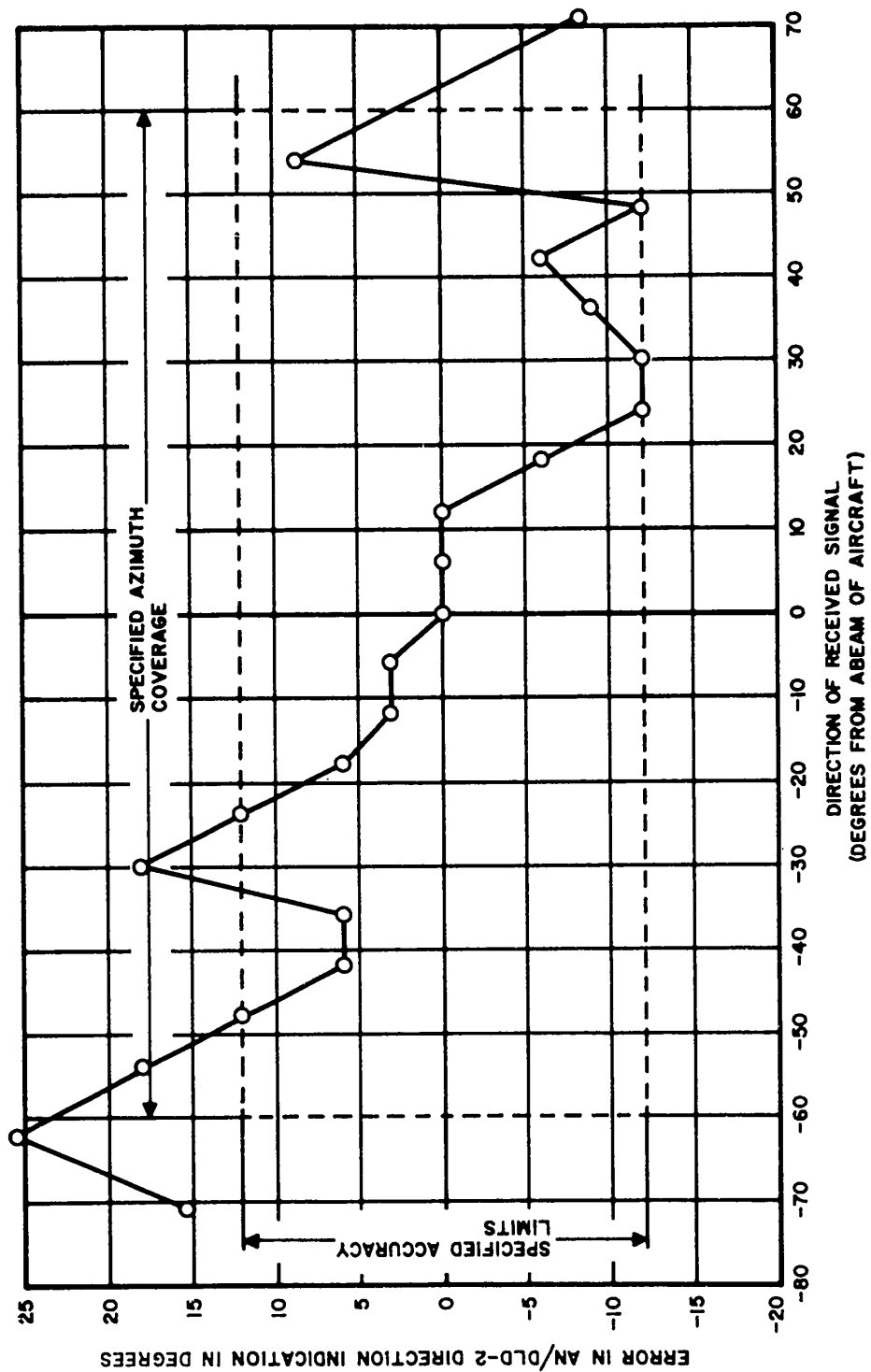


B. 93.75 MC SCALED TO 750 MC

FIGURE 122.

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C. 125 MC SCALED TO 1000 MC

FIGURE 122.

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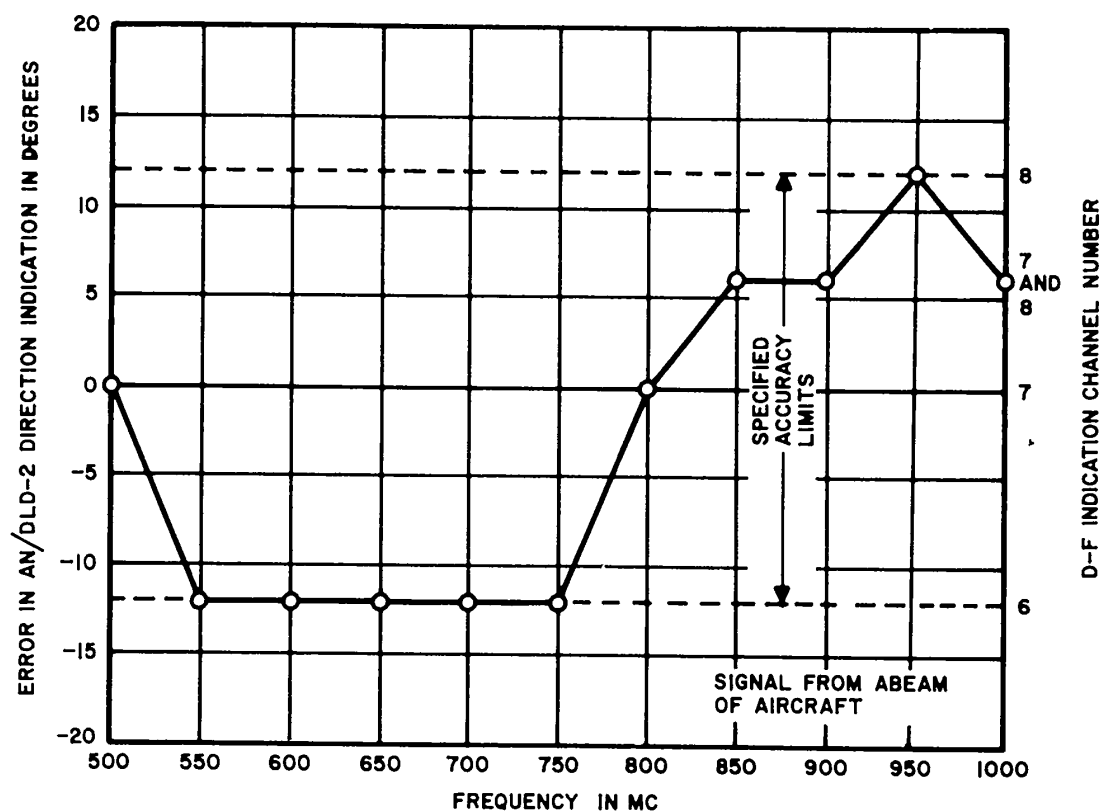
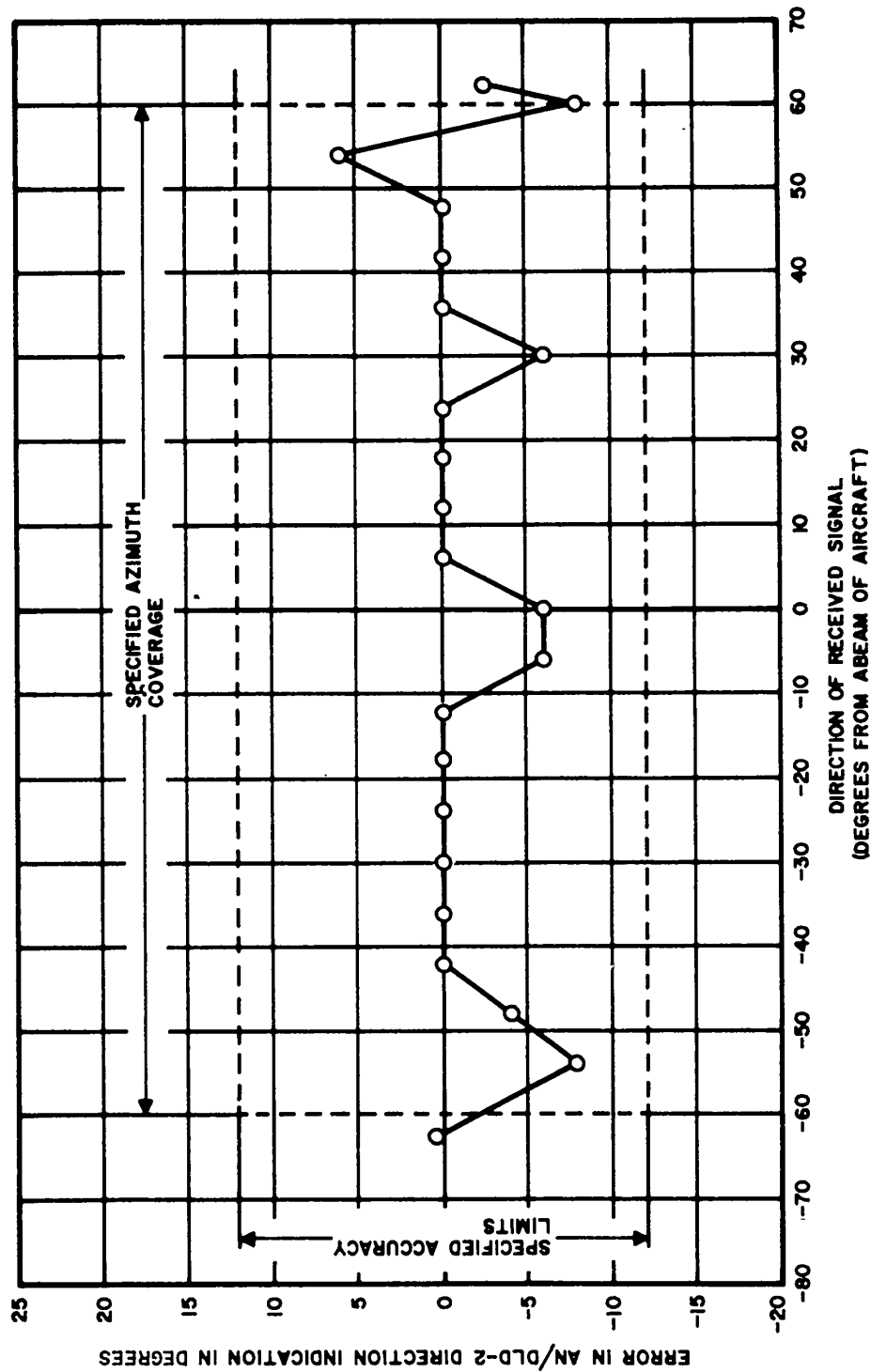


FIGURE 123. BEARING ERROR OF FEASIBILITY-DEMONSTRATION SYSTEM VS FREQUENCY IN SUB-BAND 5

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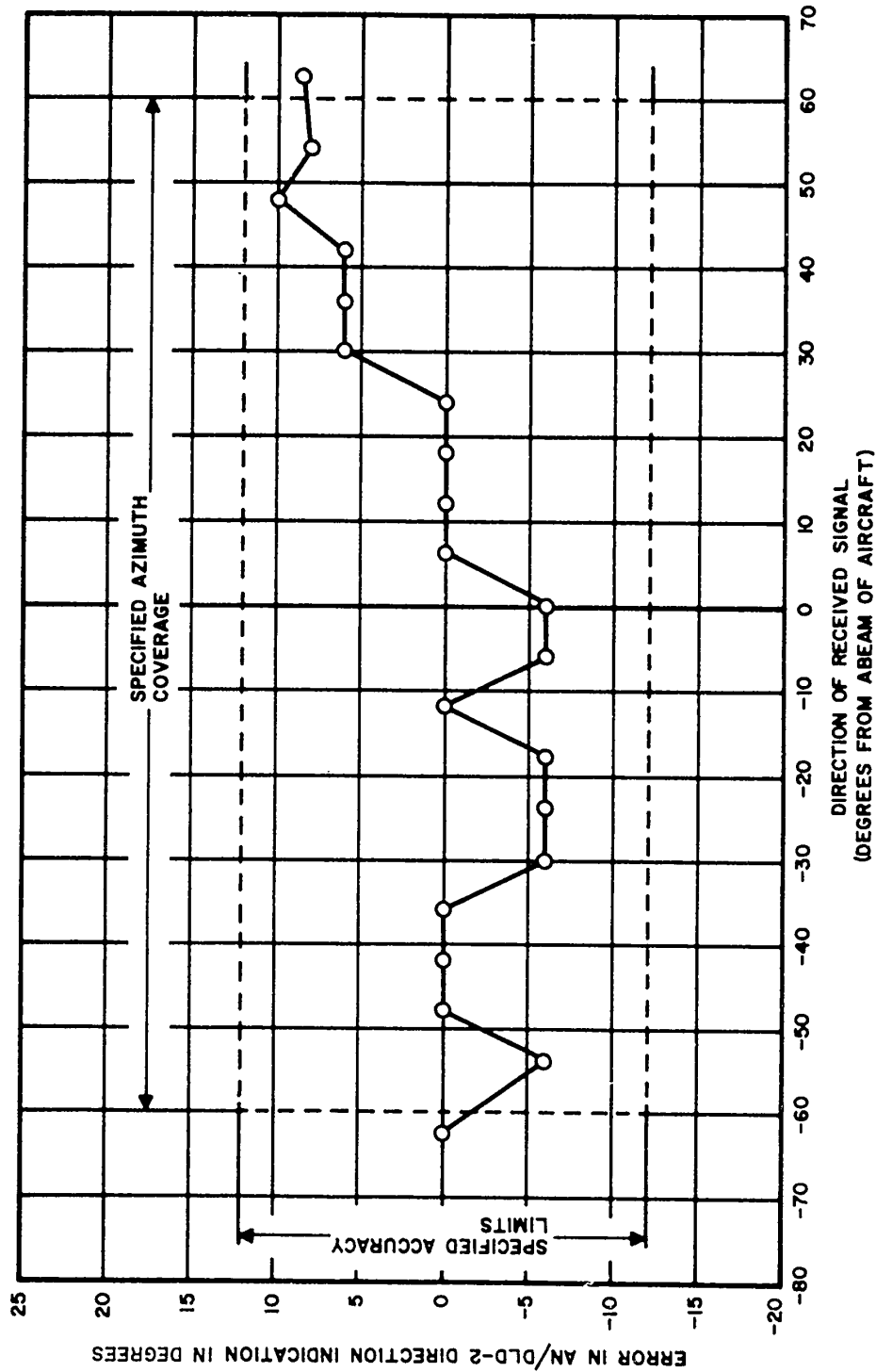


A. 550 MC

FIGURE 124. BEARING ERROR OF FEASIBILITY-DEMONSTRATION SYSTEM VS BEARING IN SUB-BAND 5

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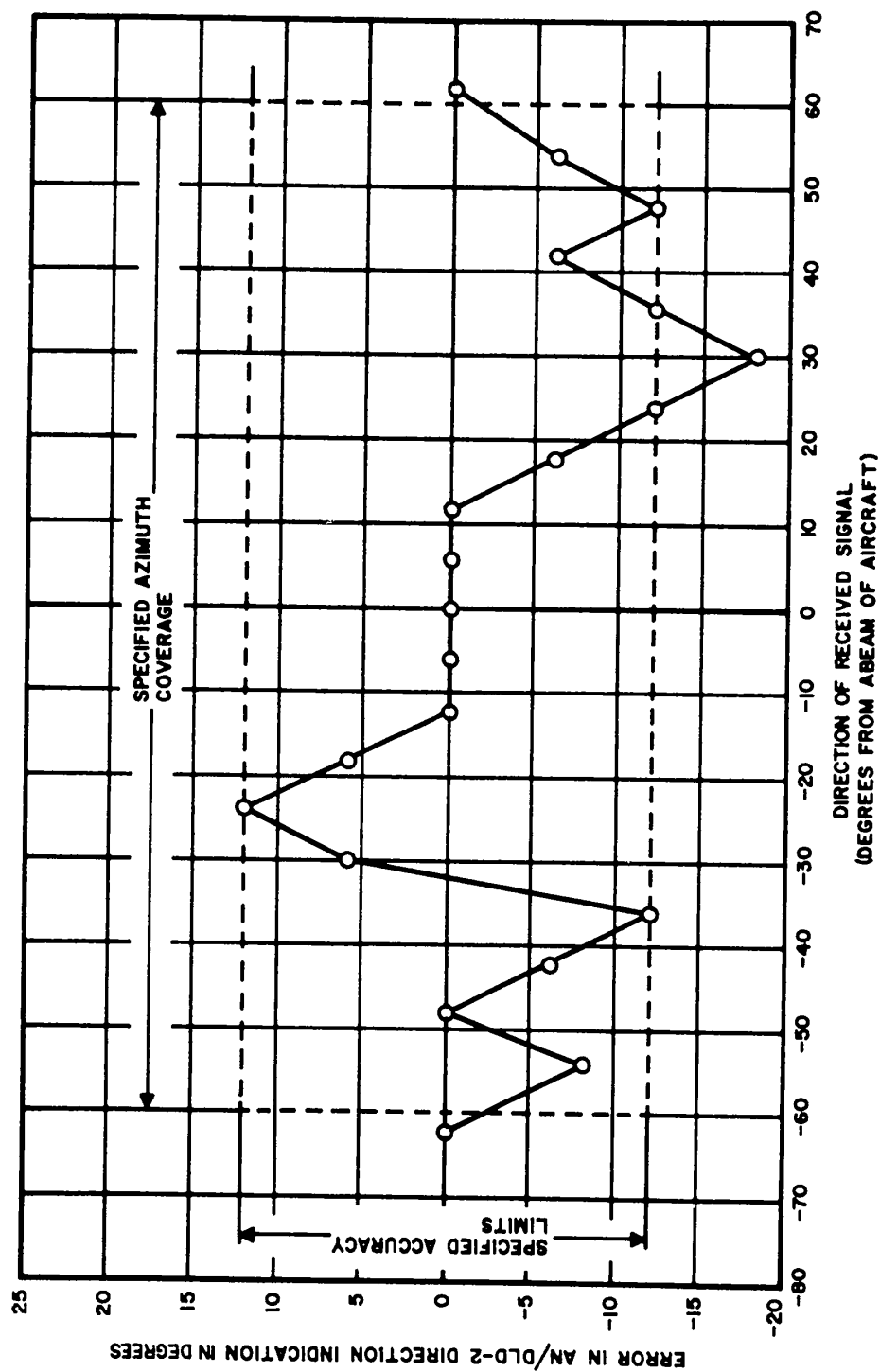
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B. 750 MC

FIGURE 124.

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C. 1000 MC

FIGURE 124.

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NOTE :
 ACCURACY DEPENDS ON BANDWIDTH.
 FILTERS USED WERE DESIGNED FOR
 AN AVERAGE ERROR (AS FUNCTION
 OF SIGNAL LEVEL) OF $\pm 2\%$.

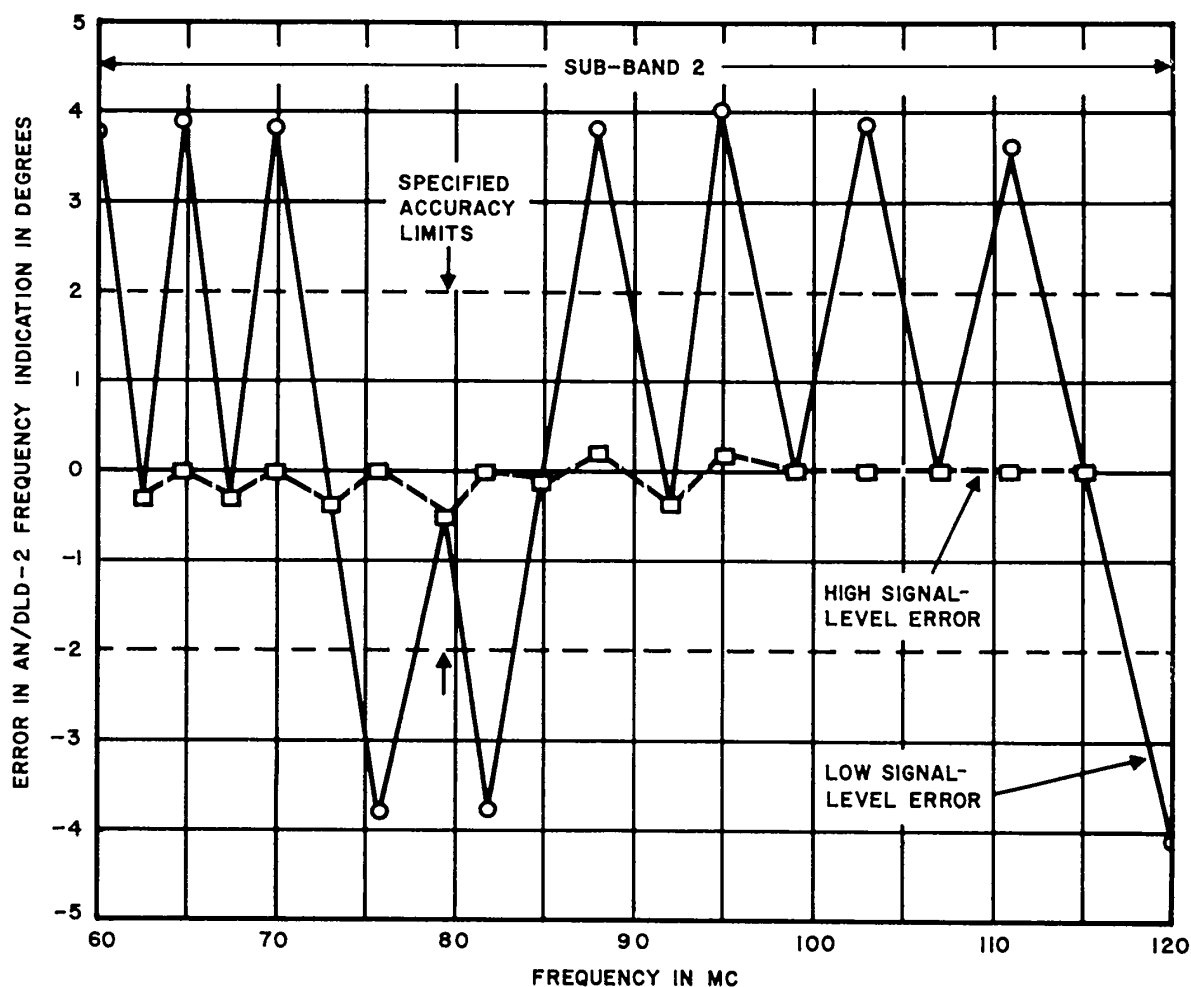


FIGURE 125. FREQUENCY ERROR VS FREQUENCY AT
 TWO POWER LEVELS IN SUB-BAND 2

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NOTE:
 ACCURACY DEPENDS ON FILTER BANDWIDTH.
 FILTERS USED WERE DESIGNED FOR AN
 AVERAGE ERROR (AS FUNCTION OF SIGNAL
 LEVEL) OF $\pm 2\%$.

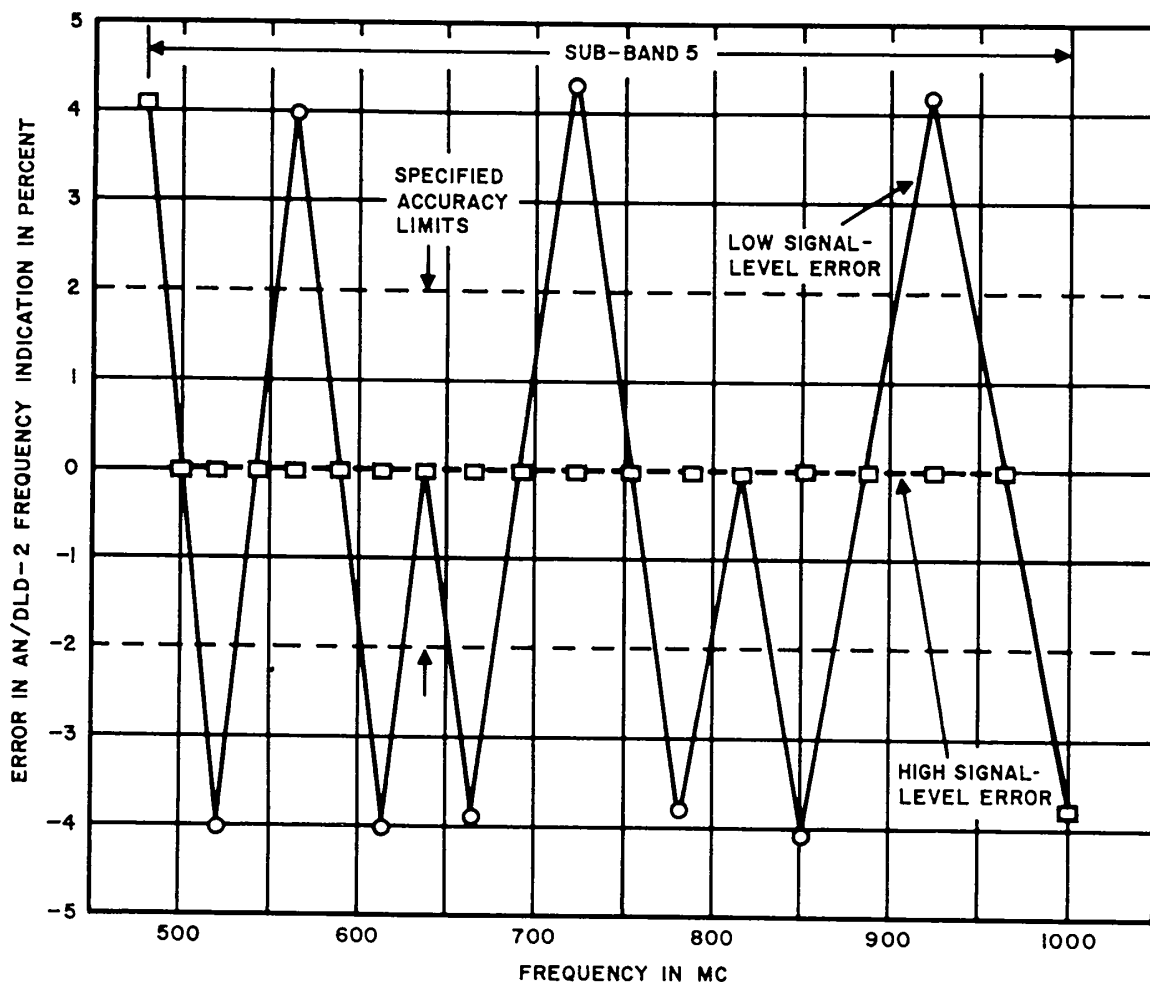


FIGURE 126. FREQUENCY ERROR VS FREQUENCY AT
 TWO POWER LEVELS IN SUB-BAND 5

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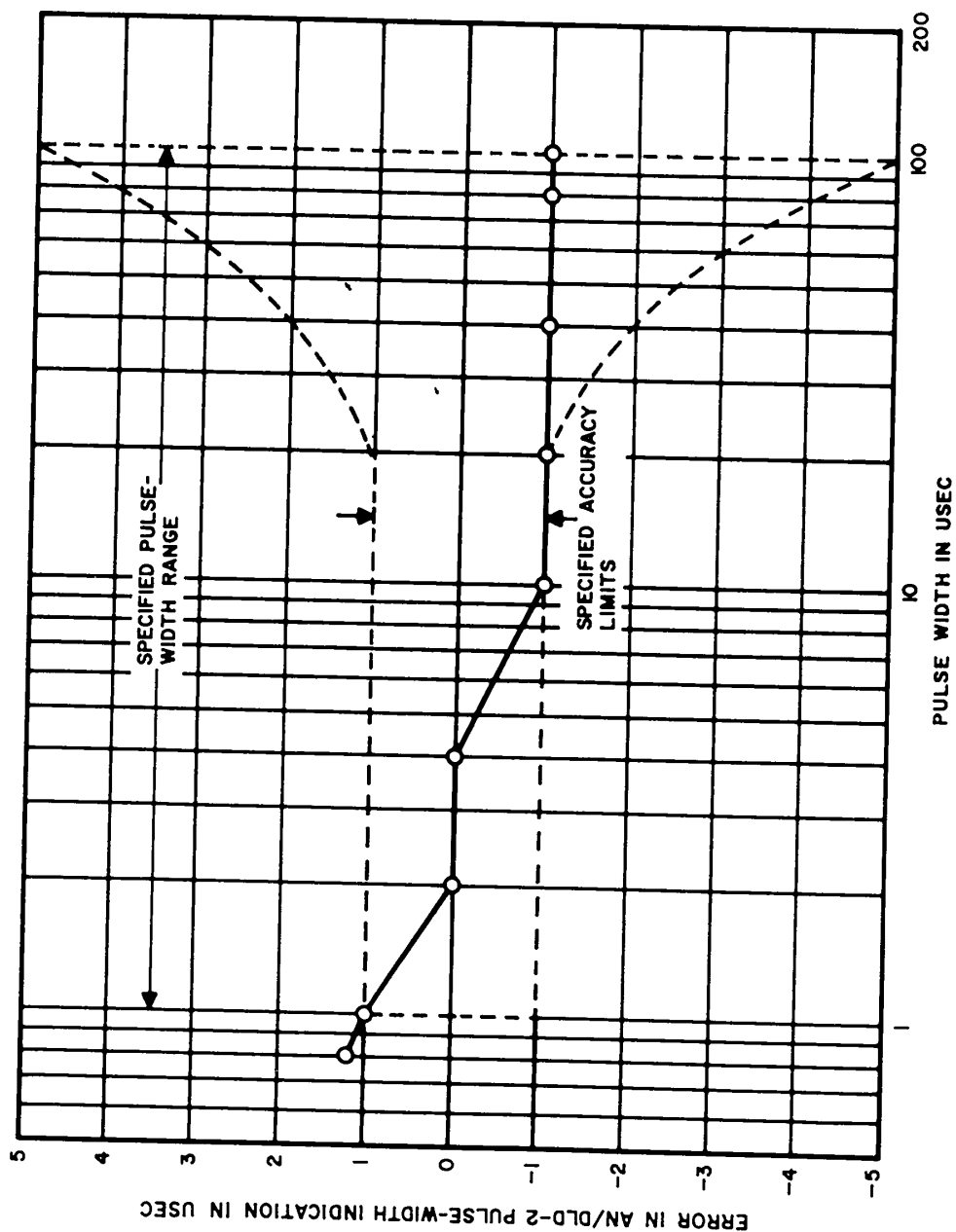


FIGURE 127. PULSE-WIDTH ERROR VS PULSE WIDTH

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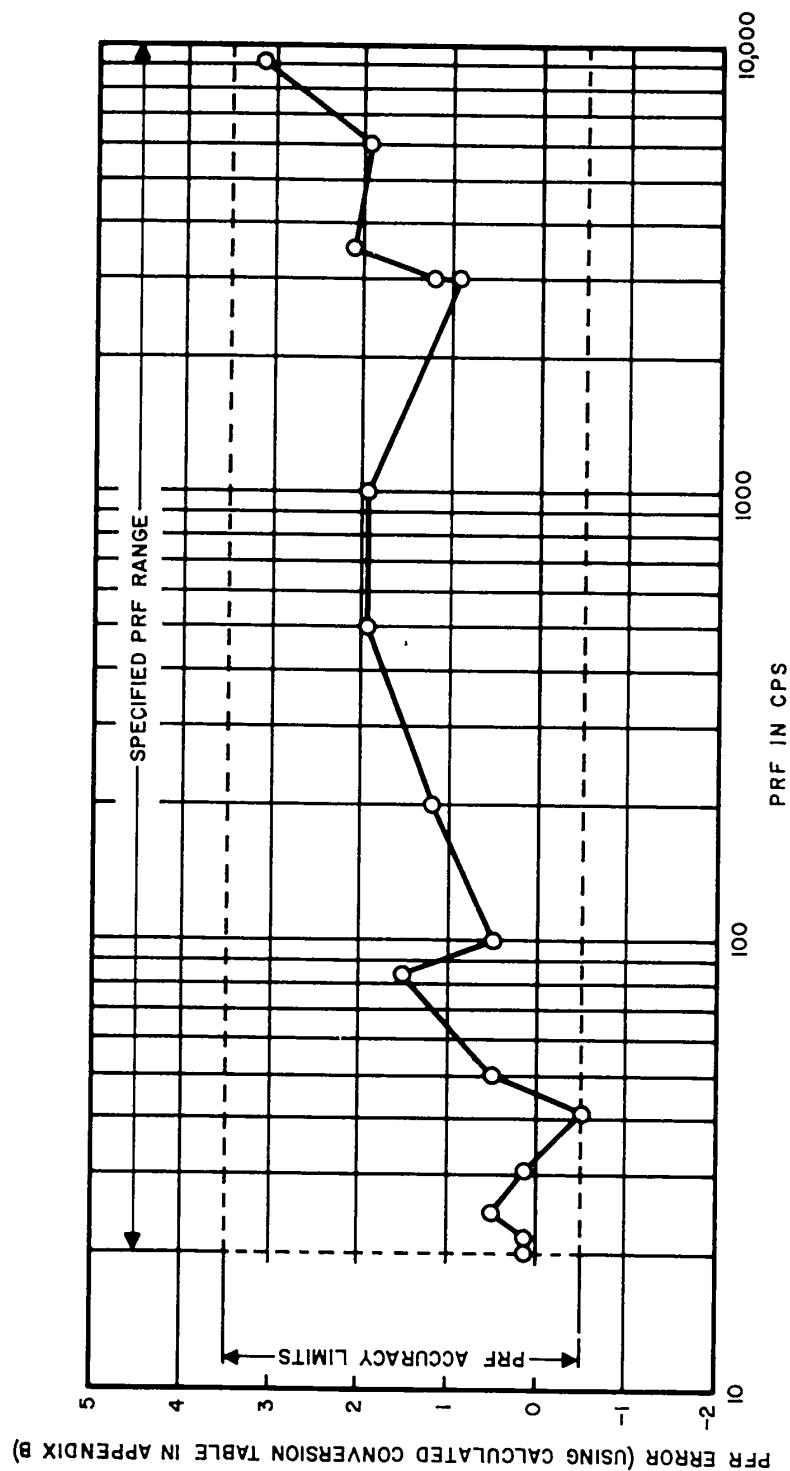


FIGURE 128. PRF ERROR VS PRF

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APPENDIX A

DESIGN OF SERIAL COMPUTER

A-1 For completeness, we are including a brief discussion of a serial computer. The design of the serial computer was carried to the stage of detailed block diagrams, wherein each block represented an individual building-block type of circuit, such as a flip-flop or a gate. This stage of design sufficed for comparison with a parallel computer that was based upon those used in the AN/DLD-1(XA-1) and earlier ferret systems designed by AIL. Since this design has not been chosen for further development, only a general block diagram (Figure A-1) is presented in this report.

A-2 The computer system operates in a series-parallel mode. The frequency and direction data are read in parallel into the 12-bit input buffer transistor shift register. These data are then shifted serially and compared with data stored in all five frequency and direction storage registers. Three conditions are possible:

1. The frequency and direction data in the input buffer do not match stored data in any of the five frequency and direction storage registers. Under this condition, the frequency and direction data are stored in an unoccupied frequency and direction storage register, and the corresponding PRF counter is energized.
2. The frequency and direction data in the input buffer do not match stored information in any of the five frequency and direction storage registers because they are occupied by other data. Under this condition, the data in the input buffer are disregarded.
3. The frequency and direction data in the input buffer are identical with the data stored in one of the five frequency and direction storage registers.

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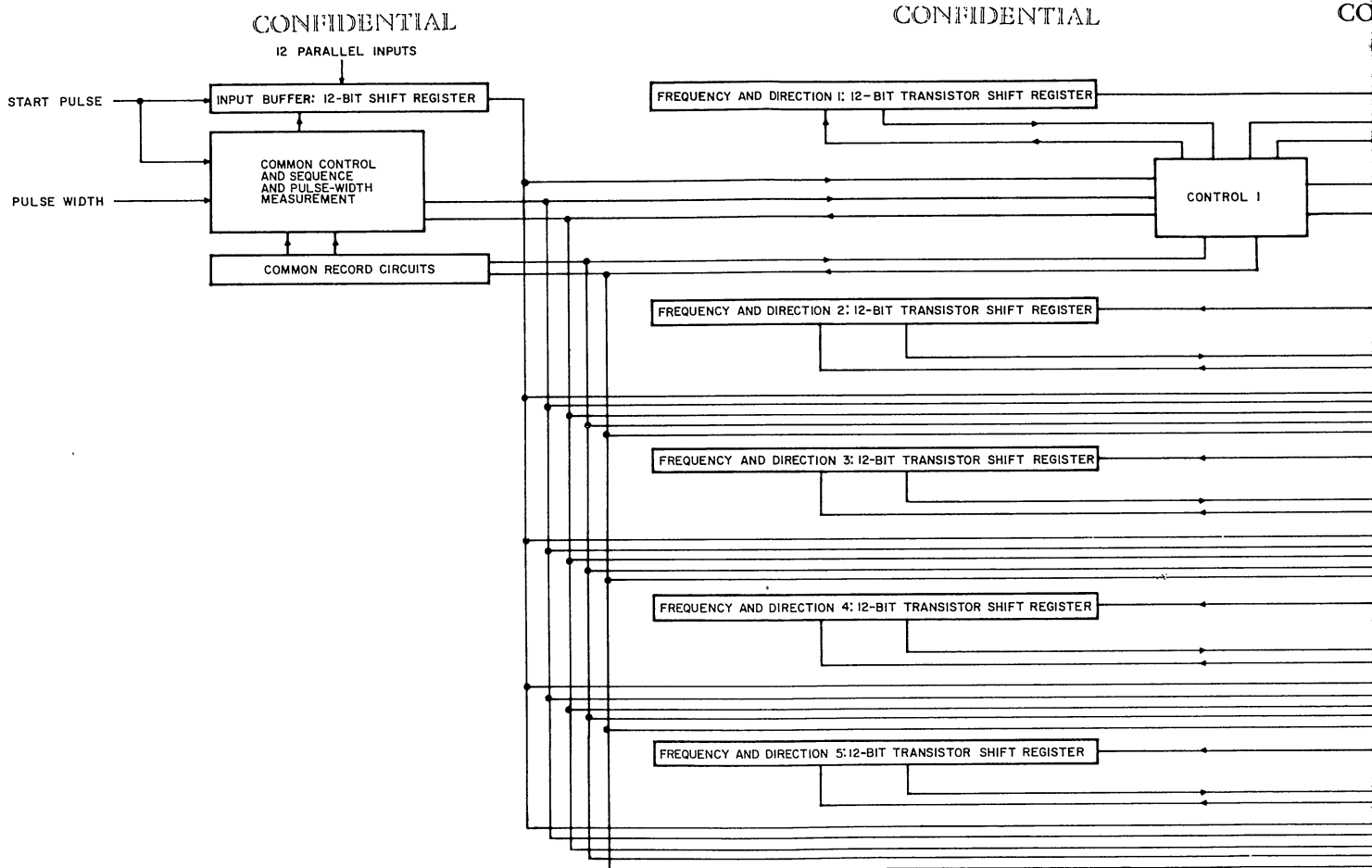
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If the intercept has been verified at the assigned time for recording the data in a pertinent channel, the channel is locked out from the remainder of the system, and the data are shifted out and recorded on magnetic tape. Data bits are shifted out in groups, each group containing the required number of bits to fill one line on the tape. If necessary, the last group of an intercept word is filled out with zeros as extra bits. When the data have been recorded, the channel is cleared and returned to its original condition as available storage. If the intercept has not been verified at the assigned time for reading the data, the channel is not locked out from the remainder of the system, and zeros are recorded on the tape for this word.

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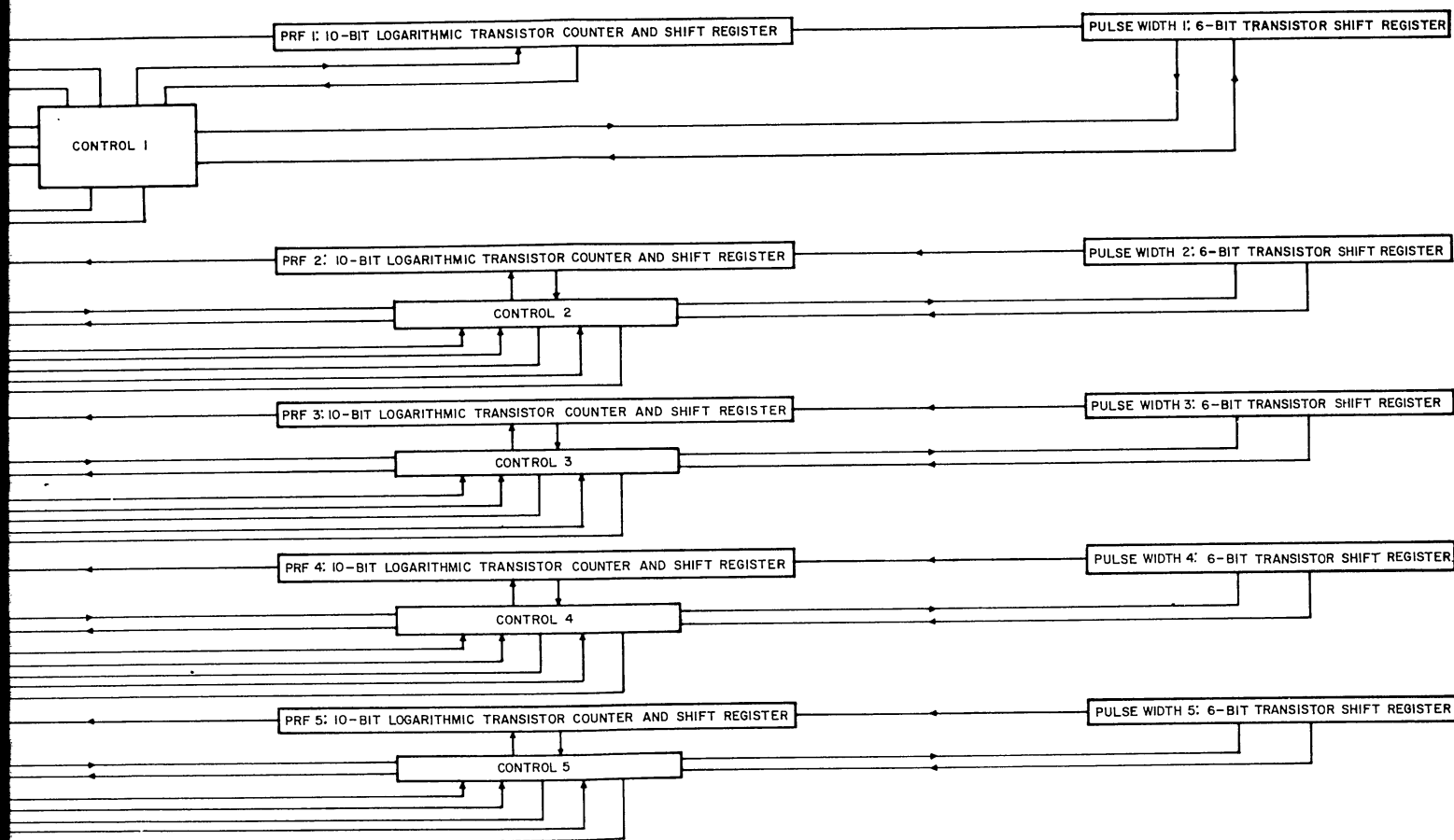


FIGURE A-1. BLOCK DIAGRAM OF SERIAL COMPUTER

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FIGURE A-1
A-5

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APPENDIX B

PRF COUNTER CALIBRATION CHART

	<u>Indication</u>	<u>PRF (pps)</u>
Where N = Counter Indication	53	10,000
For N < 258	54	9,804
PRF = $1/(N - 3)2$	55	9,615
For 257 < N < 546	56	9,434
PRF = $1/504 + (N - 258)8$	57	9,259
For 545 < N < 834	58	9,091
PRF = $1/2800 + (N - 544)32$	59	8,929
For 833 < N < 914	60	8,772
PRF = $1/12,016 + (N - 833)128$	61	8,621
For N > 913	62	8,475
PRF = $1/22,178 + (N - 912)512$	63	8,333
	64	8,197
	65	8,065
	66	7,937
	67	7,813
	68	7,692
	69	7,576
	70	7,463
	71	7,353
	72	7,246
	73	7,143
	74	7,042
	75	6,944
	76	6,849
	77	6,757
	78	6,667
	79	6,579
	80	6,494
	81	6,410
	82	6,329
	83	6,250
	84	6,173
	85	6,098
	86	6,024
	87	5,952
	88	5,882
	89	5,814
	90	5,747

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<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
91	5,682	131	3,906
92	5,618	132	3,876
93	5,556	133	3,846
94	5,495	134	3,817
95	5,435	135	3,788
96	5,376	136	3,759
97	5,319	137	3,731
98	5,263	138	3,704
99	5,208	139	3,676
100	5,155	140	3,650
101	5,100	141	3,623
102	5,051	142	3,597
103	5,000	143	3,571
104	4,950	144	3,546
105	4,902	145	3,521
106	4,854	146	3,497
107	4,808	147	3,472
108	4,762	148	3,448
109	4,718	149	3,425
110	4,673	150	3,401
111	4,630	151	3,378
112	4,587	152	3,356
113	4,545	153	3,333
114	4,505	154	3,311
115	4,464	155	3,289
116	4,425	156	3,268
117	4,386	157	3,247
118	4,348	158	3,226
119	4,310	159	3,205
120	4,274	160	3,185
121	4,237	161	3,165
122	4,202	162	3,145
123	4,167	163	3,125
124	4,132	164	3,106
125	4,098	165	3,086
126	4,065	166	3,067
127	4,032	167	3,048
128	4,000	168	3,030
129	3,968	169	3,012
130	3,937	170	2,994

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<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
171	2,976	211	2,404
172	2,959	212	2,392
173	2,941	213	2,381
174	2,924	214	2,370
175	2,907	215	2,358
176	2,890	216	2,347
177	2,874	217	2,336
178	2,857	218	2,325
179	2,841	219	2,315
180	2,825	220	2,304
181	2,809	221	2,294
182	2,793	222	2,283
183	2,778	223	2,273
184	2,762	224	2,262
185	2,747	225	2,252
186	2,732	226	2,242
187	2,717	227	2,232
188	2,703	228	2,222
189	2,688	229	2,212
190	2,674	230	2,203
191	2,660	231	2,193
192	2,646	232	2,183
193	2,632	233	2,174
194	2,618	234	2,165
195	2,604	235	2,155
196	2,591	236	2,146
197	2,577	237	2,137
198	2,564	238	2,128
199	2,551	239	2,119
200	2,538	240	2,110
201	2,525	241	2,101
202	2,513	242	2,092
203	2,500	243	2,083
204	2,488	244	2,075
205	2,475	245	2,066
206	2,463	246	2,058
207	2,451	247	2,049
208	2,439	248	2,041
209	2,427	249	2,033
210	2,415	250	2,024

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<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
251	2,016	291	1,302
252	2,008	292	1,289
253	2,000	293	1,276
254	1,992	294	1,263
255	1,984	295	1,250
256	1,976	296	1,238
257	1,969	297	1,225
258	1,984	298	1,214
259	1,953	299	1,202
260	1,923	300	1,190
261	1,894	301	1,179
262	1,866	302	1,168
263	1,838	303	1,157
264	1,812	304	1,147
265	1,786	305	1,136
266	1,761	306	1,126
267	1,736	307	1,116
268	1,712	308	1,106
269	1,689	309	1,096
270	1,667	310	1,087
271	1,645	311	1,078
272	1,623	312	1,068
273	1,603	313	1,059
274	1,582	314	1,050
275	1,563	315	1,042
276	1,543	316	1,033
277	1,524	317	1,025
278	1,506	318	1,016
279	1,488	319	1,008
280	1,471	320	1,000
281	1,453	321	992
282	1,437	322	984
283	1,420	323	977
284	1,404	324	969
285	1,389	325	962
286	1,374	326	954
287	1,359	327	947
288	1,344	328	940
289	1,330	329	933
290	1,316	330	926

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<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
331	919	371	710
332	912	372	706
333	906	373	702
334	899	374	698
335	893	375	694
336	887	376	691
337	880	377	687
338	874	378	683
339	868	379	679
340	862	380	676
341	856	381	672
342	850	382	668
343	845	383	665
344	839	384	661
345	833	385	658
346	828	386	654
347	822	387	651
348	817	388	648
349	812	389	644
350	806	390	641
351	801	391	638
352	796	392	635
353	791	393	631
354	786	394	628
355	781	395	625
356	776	396	622
357	772	397	619
358	767	398	616
359	762	399	613
360	758	400	610
361	753	401	607
362	749	402	604
363	744	403	601
364	740	404	598
365	735	405	595
366	731	406	592
367	727	407	590
368	723	408	587
369	718	409	584
370	714	410	581

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(411-490)

<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
411	579	451	488
412	576	452	486
413	573	453	484
414	571	454	483
415	568	455	481
416	566	456	479
417	563	457	477
418	561	458	475
419	558	459	473
420	556	460	472
421	553	461	470
422	551	462	468
423	548	463	466
424	546	464	465
425	543	465	463
426	541	466	461
427	539	467	460
428	536	468	458
429	534	469	456
430	532	470	455
431	530	471	453
432	527	472	451
433	525	473	450
434	523	474	488
435	521	475	446
436	519	476	444
437	517	477	443
438	514	478	442
439	512	479	440
440	510	480	439
441	508	481	437
442	506	482	436
443	504	483	434
444	502	484	433
445	500	485	431
446	498	486	430
447	496	487	428
448	494	488	427
449	492	489	425
450	490	490	424

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(491-570)

<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
491	422	531	372
492	421	532	371
493	419	533	370
494	418	534	369
495	417	535	368
496	415	536	367
497	414	537	365
498	413	538	364
499	411	539	363
500	410	540	362
501	408	541	361
502	407	542	360
503	406	543	359
504	405	544	358
505	403	545	357
506	402	546	349
507	401	547	345
508	399	548	342
509	398	549	338
510	397	550	334
511	396	551	331
512	394	552	327
513	393	553	324
514	392	554	321
515	391	555	317
516	389	556	314
517	388	557	311
518	387	558	308
519	386	559	305
520	385	560	302
521	383	561	299
522	382	562	296
523	381	563	293
524	380	564	291
525	379	565	288
526	378	566	285
527	377	567	283
528	375	568	279
529	374	569	277
530	373	570	275

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(571-650)

<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
571	272	611	202
572	270	612	201
573	268	613	199.7
574	265	614	198.4
575	263	615	197.2
576	261	616	195.9
577	259	617	194.7
578	257	618	193.5
579	254	619	192.3
580	252	620	191.1
581	250	621	190.0
582	249	622	188.8
583	247	623	187.7
584	245	624	186.6
585	243	625	185.4
586	241	626	184.4
587	239	627	183.3
588	237	628	182.2
589	235	629	181.2
590	234	630	180.1
591	232	631	179.1
592	230	632	178.1
593	228	633	177.1
594	227	634	176.1
595	225	635	175.1
596	224	636	174.1
597	222	637	173.1
598	220	638	172.2
599	218	639	171.2
600	217	640	170.4
601	215	641	169.4
602	214	642	168.5
603	212	643	167.6
604	211	644	166.7
605	210	645	165.8
606	208	646	164.9
607	207	647	164.0
608	206	648	163.2
609	204	649	162.3
610	203	650	161.5

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(651-730)

<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
651	160.7	691	133.8
652	160.0	692	132.7
653	159.0	693	132.1
654	158.2	694	131.6
655	157.4	695	131.0
656	156.6	696	130.5
657	155.9	697	129.9
658	155.1	698	129.4
659	154.3	699	128.9
660	153.6	700	128.3
661	152.8	701	127.8
662	152.1	702	127.3
663	151.3	703	126.8
664	150.6	704	126.3
665	149.9	705	125.8
666	149.2	706	125.3
667	148.5	707	124.8
668	147.8	708	124.3
669	147.1	709	123.8
670	146.4	710	123.3
671	145.7	711	122.8
672	145.0	712	122.3
673	144.3	713	121.8
674	143.7	714	121.4
675	143.0	715	120.9
676	142.4	716	120.4
677	141.7	717	120.0
678	141.1	718	119.5
679	140.4	719	119.1
680	139.8	720	118.6
681	139.2	721	118.1
682	138.6	722	117.7
683	138.0	723	117.3
684	137.4	724	116.8
685	136.8	725	116.4
686	136.2	726	116.0
687	135.6	727	115.5
688	135.0	728	115.1
689	134.4	729	114.7
690	133.8	730	114.3

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(731-810)

<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
731	113.8	771	99.4
732	113.4	772	99.0
733	113.0	773	98.7
734	112.6	774	98.4
735	112.2	775	98.1
736	111.8	776	97.8
737	111.4	777	97.5
738	111.0	778	97.2
739	110.6	779	96.9
740	110.2	780	96.6
741	109.8	781	96.3
742	109.5	782	96.0
743	109.1	783	95.7
744	108.7	784	95.4
745	108.3	785	95.1
746	107.9	786	94.8
747	107.6	787	94.6
748	107.2	788	94.3
749	106.8	789	94.0
750	106.5	790	93.7
751	106.1	791	93.4
752	105.8	792	93.1
753	105.4	793	92.9
754	105.0	794	92.6
755	104.7	795	92.3
756	104.3	796	92.0
757	104.0	797	91.8
758	103.6	798	91.5
759	103.3	799	91.2
760	103.0	800	91.0
761	102.6	801	90.7
762	102.3	802	90.4
763	102.0	803	90.2
764	101.6	804	89.9
765	101.3	805	89.7
766	101.0	806	89.4
767	100.6	807	89.2
768	100.3	808	88.9
769	100.0	809	88.7
770	99.7	810	88.4

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(81.1-890)

<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
811	88.2	851	69.8
812	87.9	852	69.2
813	87.7	853	68.6
814	87.4	854	68.0
815	87.2	855	67.4
816	86.9	856	66.8
817	86.7	857	66.3
818	86.4	858	65.7
819	86.2	859	65.2
820	86.0	860	64.6
821	85.7	861	64.1
822	85.5	862	63.6
823	85.3	863	63.1
824	85.0	864	62.6
825	84.8	865	62.1
826	84.6	866	61.6
827	84.3	867	61.1
828	84.1	868	60.6
829	83.9	869	60.0
830	83.7	870	59.7
831	83.4	871	59.2
832	83.2	872	58.8
833	82.8	873	58.4
334	82.3	874	57.9
835	81.5	875	57.5
836	80.6	876	57.1
837	79.8	877	56.7
838	79.0	878	56.3
839	78.2	879	55.9
840	77.4	880	55.5
841	76.7	881	55.1
842	75.9	882	54.7
843	75.2	883	54.3
844	74.5	884	53.9
845	73.8	885	53.6
846	73.1	886	53.2
847	72.4	887	52.8
848	71.8	888	52.5
849	71.1	889	52.1
850	70.5	890	51.8

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(891-970)

<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
891	51.4	931	37.0
892	51.1	932	36.6
893	50.8	933	36.3
894	50.4	934	36.0
895	50.1	935	35.6
896	49.8	936	35.3
897	49.5	937	35.0
898	49.2	938	34.7
899	48.9	939	34.4
900	48.6	940	34.1
901	48.3	941	33.8
902	48.0	942	33.5
903	47.7	943	33.2
904	47.4	944	33.0
905	47.1	945	32.7
906	46.8	946	32.4
907	46.5	947	32.1
908	46.3	948	31.9
909	46.0	949	31.6
910	45.7	950	31.3
911	45.5	951	31.1
912	45.2	952	30.8
913	44.9	953	30.6
914	44.1	954	30.4
915	43.6	955	30.1
916	43.1	956	29.9
917	42.6	957	29.7
918	42.2	958	29.5
919	41.7	959	29.2
920	41.3	960	29.0
921	40.9	961	28.8
922	40.4	962	28.6
923	40.0	963	28.4
924	39.6	964	28.2
925	39.2	965	28.0
926	38.8	966	27.8
927	38.4	967	27.6
928	38.1	968	27.4
929	37.7	969	27.2
930	37.3	970	27.0

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(971-1023)

<u>Indication</u>	<u>PRF (pps)</u>	<u>Indication</u>	<u>PRF (pps)</u>
971	26.8	1011	21.2
972	26.6	1012	20.9
973	26.5	1013	20.8
974	26.3	1014	20.7
975	26.1	1015	20.6
976	25.9	1016	20.5
977	25.8	1017	20.4
978	25.6	1018	20.3
979	25.4	1019	20.2
980	25.3	1020	20.1
981	25.1	1021	20.0
982	24.9	1022	19.9
983	24.8	1023	19.8
984	24.6		
985	24.5		
986	24.3		
987	24.2		
988	24.0		
989	23.9		
990	23.7		
991	23.6		
992	23.4		
993	23.3		
994	23.2		
995	23.0		
996	22.9		
997	22.8		
998	22.6		
999	22.5		
1000	22.4		
1001	22.2		
1002	22.1		
1003	22.0		
1004	21.9		
1005	21.7		
1006	21.6		
1007	21.5		
1008	21.4		
1009	21.3		
1010	21.2		

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STAT

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Translation from the German of newspaper article from the July 31, 1958 issue of the ST. GALLER TAGBLATT.

Service of Red Espionage

Technology in the Espionage Service

Since the discovery of some telephone-spies in West Berlin, who tapped the telephones of American and German posts under Soviet instructions, and especially since the arrest a month later of a Dutch businessman in West Berlin, the western counter-intelligence offices have seized upon feverish new security measures. The arrested Dutchman had actually furnished the East's secret services with refined mechanisms of espionage--from a bogus firm--indeed, he trafficked there about thirty cigarette-lighters with excellently-made secret cameras, twenty wrist-watches and fountain pens with tiny hidden microphones, as well as about 10,000 so-called "Cobbler's Rogues" which lent themselves perfectly to a carpenter's insertion in an false passage. (

It has been known that the Soviets have long specialized in this technical espionage since the days of the proceedings against the Communist Part by the West German tribunal. Here one of the East German defenders was observed as he listened to a discussion that the Tribunal was conducting behind closed doors in the council-room, using a telephonic apparatus in his automobile at a distance of 100 meters. The apparatus used here, which is known in the United States as "parabolic microphones" and which each great intelligence service has already had in its possession for a long time, was developed originally by ornithologists, who wanted to record the voices of especially shy birds. It can be activated by any acoustic source, like a camera, and the recording of the conversation flows over a shell in parabolic form. Conversation with one's self, held in a whispering tone, can be heard at a distance of 100 meters. Also, conversations held in a room can, using the apparatus, be heard, providing a window in the room is open, as was accidentally the case in Karlsruhe. The most effective devices of this kind can pick up conversations at a distance of 250 meters! (sic) The experts point out that with them even conversations in closed rooms are capable of being heard if ever one should succeed in smuggling a tiny transmitter into the body of a cigarette-lighter or a fountainpen. The transmitter only needs to amplify the noises enough so that they pierce weakly through the window-panes or walls (with microwaves is this easily possible) and the receiving telephonic device picks the sound-waves up there; with one of the listening devices then hooked up to a "Magnetophone", the overheard talk can then be recorded without further effort and usually played back.

In the pen-holders and wrist-watches which the Dutchman delivered to East Berlin different kinds of tiny transmitters were not built in. They held only tiny microphones, which were however tied in by a thin wire to a special receiving device which could be brought in in the vest-pocket of any possessor.

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[illegible]

This technical espionage is therefore becoming ever more refined, so that at the present a device is being developed in the U.S.A. as well as in Russia that will be able to tap telephone wires through special induction streams (currents) directed by arrangements of coils without having to actually tap the wires in the old sense of the word. It would consist in gathering various devices at central points, so that it is entirely possible to bring the entire telephone net of a city under control and to listen to and record each conversational (sic).

In the non-holders and all the rest which the Director delivered to East Berlin district office. They paid only five marks for it. The other three, which were forwarded in the third class, were received by the recipient's office. It could be brought in in the first class.

STAT

Die Technik im Dienst roter Spionage

Seit der Entdeckung einiger Telefonspione in Westberlin, die im Auftrage der Sowjets die Telephone amerikanischer und deutscher Dienststellen angezapft hatten, und insbesondere seit der vor einem Monat erfolgten Verhaftung eines holländischen Geschäftsmannes in Westberlin, werden bei der westlichen Spionageabwehrstellen lebhaft neue Sicherungsmaßnahmen getroffen. Der festgenommene Holländer hatte nämlich über eine Scheinfirma in Ostberlin die östlichen Geheimdienste mit raffinierten «Spionagerobotern» beliefert — und zwar handelte es sich dabei um 30 Feuerzeuge mit ausgezeichnet getarnten Geheimkameras, 20 Armbanduhrn und Füllfederhalter mit winzigen Geheimmikrophonen sowie um rund 10 000 sogenannte «Schusterösen», die sich glänzend zum Einsetzen von Paßbildern in gefälschte Pässe eignen.

Daß sich die Sowjets schon seit geraumer Zeit auf diese technische Spionage spezialisiert haben, ist seit den Tagen des Prozesses gegen die Kommunistische Partei vor dem Westdeutschen Bundesgerichtshof in Karlsruhe bekannt. Hier wurde einer der ostdeutschen Verteidiger dabei beobachtet, wie er mit einem Telehorchgerät von seinem Auto aus über eine Entfernung von 100 Metern Gespräche abhörte, die das Gericht hinter verschlossenen Türen im Beratungszimmer führte. Das hierzu benutzte Gerät, das in den Vereinigten Staaten unter der Bezeichnung «parabolic microphone» bekannt ist und das jeder größere Geheimdienst bereits seit längerer Zeit besitzt, wurde

ursprünglich von Ornithologen entwickelt, um die Stimmen besonders scheuer Vögel festzuhalten. Es kann auf jede beliebige akustische Quelle wie eine Kamera eingerichtet werden, und die Aufnahme des Gesprächs erfolgt über eine Muschel in parabolischer Form. Selbst Gespräche, die im Flüsterton geführt werden, kann man mit ihm über eine Entfernung bis zu 100 Meter mithören. Auch Unterhaltungen, die in einem Zimmer geführt werden, sind mit dem Gerät zu belauschen, sofern ein Fenster im Raume offensteht — und in Karlsruhe war das zufällig der Fall. Die leistungsfähigsten Geräte dieser Art können Gespräche sogar bis zu 250 Meter Entfernung aufnehmen! Experten weisen darauf hin, daß mit ihnen sogar Unterhaltungen in verschlossenen Räumen abgehört werden könnten, wenn es im Einzelfall gelingen sollte, in diese einen winzigen Sender in der Größe eines Feuerzeuges oder eines Füllfederhalters zu schmuggeln. Der Sender braucht dabei die Geräusche nur so weit zu verstärken, daß sie schwach durch die Fensterscheiben oder Wände dringen (mit Megawellen ist das durchaus möglich), und das betreffende Telehorchgerät nimmt die Schallwellen von dort auf; mit einem dem Horchgerät dann angeschlossenen Magnetophon kann das Abgehörte ohne weiteres konserviert und beliebig oft abgespielt werden.

In den Füllfederhaltern und den Armbanduhrn, die der Holländer nach Ostberlin geliefert hatte, waren derartige Kleinsender nicht eingebaut. Sie enthielten zwar nur Mikrophone, die aber durch einen dünnen Draht mit einer Spezialaufnahmeapparatur verbunden waren, die sich leicht in der Westentasche des jeweiligen Besitzers unterbringen läßt. Nach westlichen Informationen waren die Konstruktionen solcher winziger Mikrophone und insbesondere solcher Aufnahmegeräte bisher im gesamten Ostblock unbekannt. Mit Hilfe derartiger Sender, die nunmehr

gegen sich aufbringen würden. Darauf ist die Frage: «Heißt dies, daß das Informations einseitig, gelenkt und dirigiert sein soll?» Die kürzlichen Personalverschiebungen im Rat lassen? Gehört es nicht zum Wesen der Demokratie, die Diskussion zu gewährleisten, auf die Intelligenz des Bürgers zu vertrauen, der sich selbst und gegensätzlich informiert wird? Und Mayer schließt: «Ich halte im übrigen eine solche Erklärung für nutzlos. Ich mache sie nur, um nicht als Bube eines scheinbar freien Radiowesens aufzutreten, denn frei ist dieses nicht oder nicht mehr, der gewandte Vorfall zeigt das Ausmaß seiner tatsächlichen Freiheit».

Nun, man kann sich des Eindrucks nicht erwehren, Soustelle hatte durch die Zulassung einer selbst direkt gegen ihn gerichteten kurzen Bemerkung sein Pressé nur fordern und gewisse Hoffnungen auf Zerstreuung können. Aber eine solche Haltung ist schließlich einen tief verwurzelten Glauben des Informationsministers an die demokratischen Ideale zu setzen ...

auch in den Händen der Sowjets sind, kann man darüber hinaus Gespräche in einem fahrenden Auto mithören; andere wieder brauchen nur an die Wand eines Zimmers befestigt zu werden, um sofort die im Nebenzimmer geführte Unterhaltung abhören zu können.

Diese technische Spionage wird dadurch nur noch raffinierter, daß gegenwärtig sowohl in den USA als auch in Rußland ein Gerät entwickelt wird, mit dem Telefonleitungen angezapft werden können durch über besondere Spulenarrangierungen geleitete Induktionsströme — ohne sie in eigentlichen Sinne des Wortes direkt anzapfen zu müssen. Man bringt es, derartige Geräte an zentrale Punkte anzubringen, so ist es durchaus möglich, das gesamte Telefonnetz einer Stadt unter Kontrolle zu bringen und auch jedes Gespräch mithören zu können.

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